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Artificial transmission lines in television circuits

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ARTIFICIAL TRANSMISSION LINES IN TELEVISION CIRCUITS

HAROLD RAYMOND WALKER

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H. R. WALKER

ARTIFICIAL TRANSMISSION LINES

IN TELEVISION CIRCUITS

By

Harold Raymond Walker
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Submitted in partial fulfillment
of the requirements
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Certificate of Completion in Engineering Electronics

From the

United States Naval Postgraduate School

PREFACE

Many television circuits and auxiliary devices have appeared on the market recently which incorporate artificial transmission lines as a basic part of their design. The two main functions served by artificial transmission line sections are impedance matching and capacity isolation, but their use for these functions make possible extremely wide bandwidth and comparatively low noise figures when compared with tuned circuits as ordinarily used.

This thesis develops the mathematics necessary for the design of artificial transmission line sections in the resonant single frequency case and by means of transfer functions extends this to the broadband case as normally used. Much of the work on extending the operating frequency over a broadband is original, as are the transfer functions used. This work is not done in any reference known to the author, although it is done by cut and try methods in most laboratories which work with broadband amplifiers.

Because mathematics do not present a clear and immediate picture of the conditions prevailing, graphs are used to permit rapid calculation by cut and try methods on paper.

In order to limit the scope of this thesis, only L, T and pi networks are treated, but where an additional improvement results from the use of M dervied or constant K sections, this is pointed out. Since the uses of artificial transmission lines given here are essentially lossless, the case of sections with losses will not be considered.

Following the presentation of the theory involved is a survey of circuits employing this theory at the present time, plus several possible circuits developed by the author which offer desirable features to television and radar equipment.

The writer wishes to thank Professors P. E. Cooper, D. A. Stentz, and G. J. Thaler for their assistance and advice in doing the necessary laboratory work, and for their constructive criticism in the preparation of this thesis.

TABLE OF CONTENTS

LIST OF ILLUSTRATIONS

TABLE OF SYMBOLS USED

C	-	Capacitance
E_1	-	Input Voltage
E_2	-	Output Voltage
E(s)	-	Voltage a function of (s) (Laplace Transform)
fc	-	Cutoff Frequency
fr	-	Resonant Frequency
G2	-	Shunt Conductance
$I_{(n)}$	-	Loop Current
$I_{(n)}(s)$	-	Current a function of (s) (Laplace Transform)
L	-	Inductance
Q	-	Quality factor - See Definition Page 2.
R_1	-	Series Resistance
R_2	-	Shunt Resistance
R_c	-	Equivalent Series Resistance in Condenser used in place of 2 in certain formulas.
R_L	-	Load Resistance
R_p	-	Plate Resistance
(s)	-	A variable used in Laplace Transforms (jw)
t	-	Time
X_c	-	Capacitive Reactance
X_L	-	Inductive Reactance
Y	-	Admittance
Z	-	Impedance
Z_o	-	Nominal Line Impedance

CHAPTER I

BASIC THEORY OF MATCHING SECTIONS

The resonant circuit shown in figure (1) has a series resistance R_1 and a shunting conductance G_2 which shall be interchangeably referred to as R_2. If this circuit is resonant at a frequency determined by L, C and R, the relationship given in quation (1) gives the driving point impedance of the anti-resonant circuit. If no conductance G_2 is assumed, and the only loss element is R_1, then at resonance the equation (2) holds. We thus have a means of converting from series to shunt resistance. This relationship is subject to further transformation so that equation (3) results. Equations (4) and (5) are taken from transmission line theory and given here so that their similarity will be noted.

It should further be noted that equation (5) is the equation for matching two line impedances by means of a quarter wave stub. If figure (1) is redrawn as in figure (2), the same anti-resonant circuit becomes an L section which will match R_1 to R_2. The range of values over which the L section will match impedances is theoretically infinite, however, due to the practical aspects of the circuit, losses become excessive if the Q of the circuit exceeds 10, in which case it is better to use several L sections, each with a lower Q.

The relationship of equation (3) is subject to further manipulation to get equations (6) and (7). Equation (7) is that of a phase and amplitude distortionless line. For a single frequency match, the

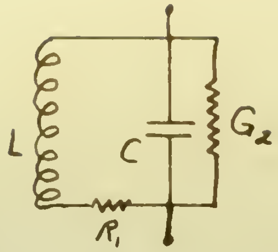


Fig. 1

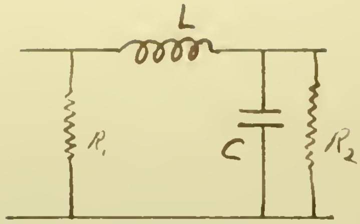


Fig. 2

$$Z_r \doteq \frac{L}{CR} \qquad (1)$$

$$R_2 = \frac{L}{CR_1} \qquad (2)$$

$$R_1 R_2 = L/C \qquad (3)$$

$$Z_0 \doteq \sqrt{L/C} \qquad (4)$$

$$\sqrt{Z_1 Z_2} = Z_0 \qquad (5)$$

$$\frac{L}{C} = \frac{\omega L}{\omega C} = X_L X_C = Z_0^2 \qquad (6)$$

$$R_2 = \frac{L}{CR_1} \qquad R_2 C = \frac{L}{R_1} = \frac{C}{G} \qquad (7)$$

$$Q = 2\pi \frac{\text{energy stored}}{\text{energy lost}} \text{ (per cycle)} \qquad (8)$$

$$\frac{R_2}{X_C} = \frac{X_L}{R_1} = Q \qquad (9) \qquad X_C = \frac{R_2}{Q} \qquad X_L = R_1 Q$$

$$\frac{R_2}{X_C} = \frac{X_L}{R_1} = Q \qquad (9)$$

$$R_2 X_C = R_2 X_L + R_1 X_C$$

Dividing thru by $X_C R_1$

$$\frac{R_2 X_C}{R_1 X_C} = \frac{R_2 X_L}{R_1 X_C} + \frac{R_1 X_C}{R_1 X_C}$$

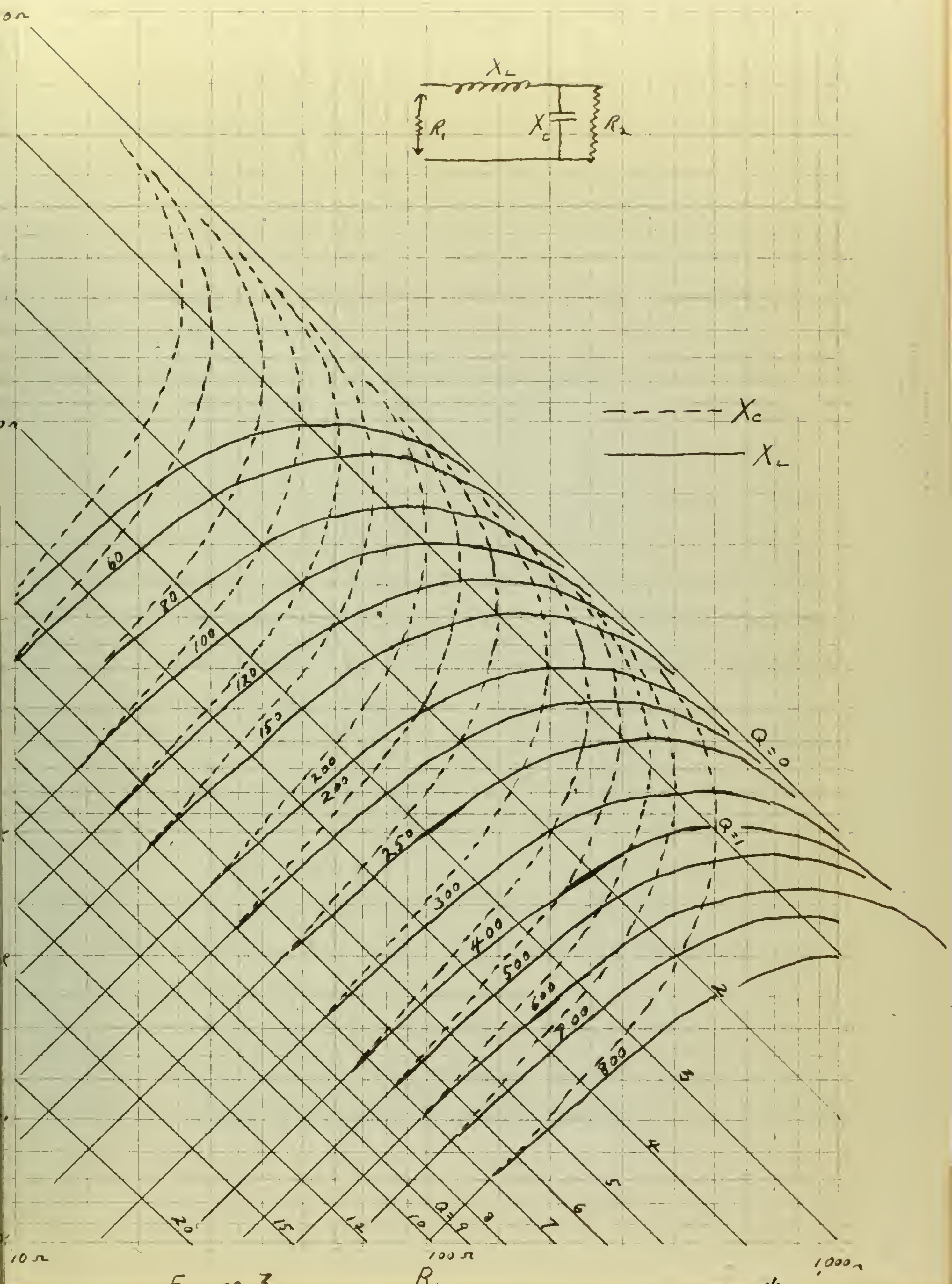
$$\frac{R_2}{R_1} = Q^2 + 1 \qquad (10)$$

apparent solution is to pick X_L, X_c and Z_o all equal. Under those conditions resonance is assumed and the input and output should be pure resistive components. For low Q matches, ie., matches where R_1 and R_2 are not widely different, X_L cannot equal X_c as will be apparent from a further investigation of resonance.

The resonant circuit has a Q defined by equation (8). Q may also be defined as the ratio of reactance to resistance as given by equation (9), and as the tangent of the phase angle of current delay or advance. Manipulating equation (9) we obtain equation (10) which is the basis of low Q matching.

Figure (3) gives a universal matching chart for L matching R_1 to R_2. It will be noted from this chart that the diagonal co-ordinate is Q, and that as Q approaches zero, the capacitive reactance required for a match is greater than the inductive reactance. The reason for this becomes apparent when the basic resonance equation is expanded. Defining resonance as the condition of unity power factor, or pure resistive components, the general resonance formula is given by equation (11). Under conditions of high Q's, R_1 and R_c are small, and the approximation, equation (12) is employed. R_c is the series equivalent of the shunt resistor R_2 which would be assumed to be a resistor in series with C. For most applications other than this formula, it is more convenient to leave it as a conductance G, which represents leakage, tube input conductance, or just a resistive load.

As an example of how Q affects L and C in the L matching circuit, let $Q = 1$. When $Q = 1$, R_2 is two times R_1 by equation (10), and



X_L
R_1
X_c
R_2
X_c
X_L
60
80
100
120
150
200
200
250
300
400
500
600
700
800
Q=0
Q=1
2
3
4
5
6
7
8
Q=9
10
12
15
20
10 Ω
100 Ω
1,000 Ω
R_1

Figure 3

similarly R_c is two times R_1. Taking the large radical in equation (11), substituting equation (13), we obtain equation (14) which indicates the actual resonant frequency is only .707 that indicated by equation (12). Also, since equation (10) states $R_2 = 2R_1$, then X_c must equal $2X_L$ if equation (9) is to be satisfied. It can be seen from figure (3) that $X_c = 2X_L$ when $Q = 1$. (15)

When a chart, such as figure (3), is not available, a convenient rule of thumb using equation (15) may be set up. To use this equation determine Q by equation (10) from the known impedances to be matched. Pick X_c from equation (9) and then pick an X_L which will equal X_c at the frequency given by equation (15). Thus for $Q = 1$, X_L would equal X_c at a frequency 1.4 times the frequency of perfect match. Failure to make the low Q reactance correction results in a loss of about 2 db in most cases, but the match then becomes reactive and may result in distortions in other parts of the circuit.

The basic L section is only a step in the complete matching process. Often it may be enough, but generally several sections are required either front to front as pi sections, end to end as a ladder network, or back to back as a T section. If R_1 is made equal to R_2, $Q = 0$ and R_2 can be replaced by another similar L section. When this is done, several times, an artificial transmission line terminated in Z_o results. Figure (4) illustrates these various combinations.

It should be noted that the artificial line is symmetrical, and thus has the same transmission characteristics, both ways. Further, any mismatch in Z_o will cause impedances to reflect from point to point and the

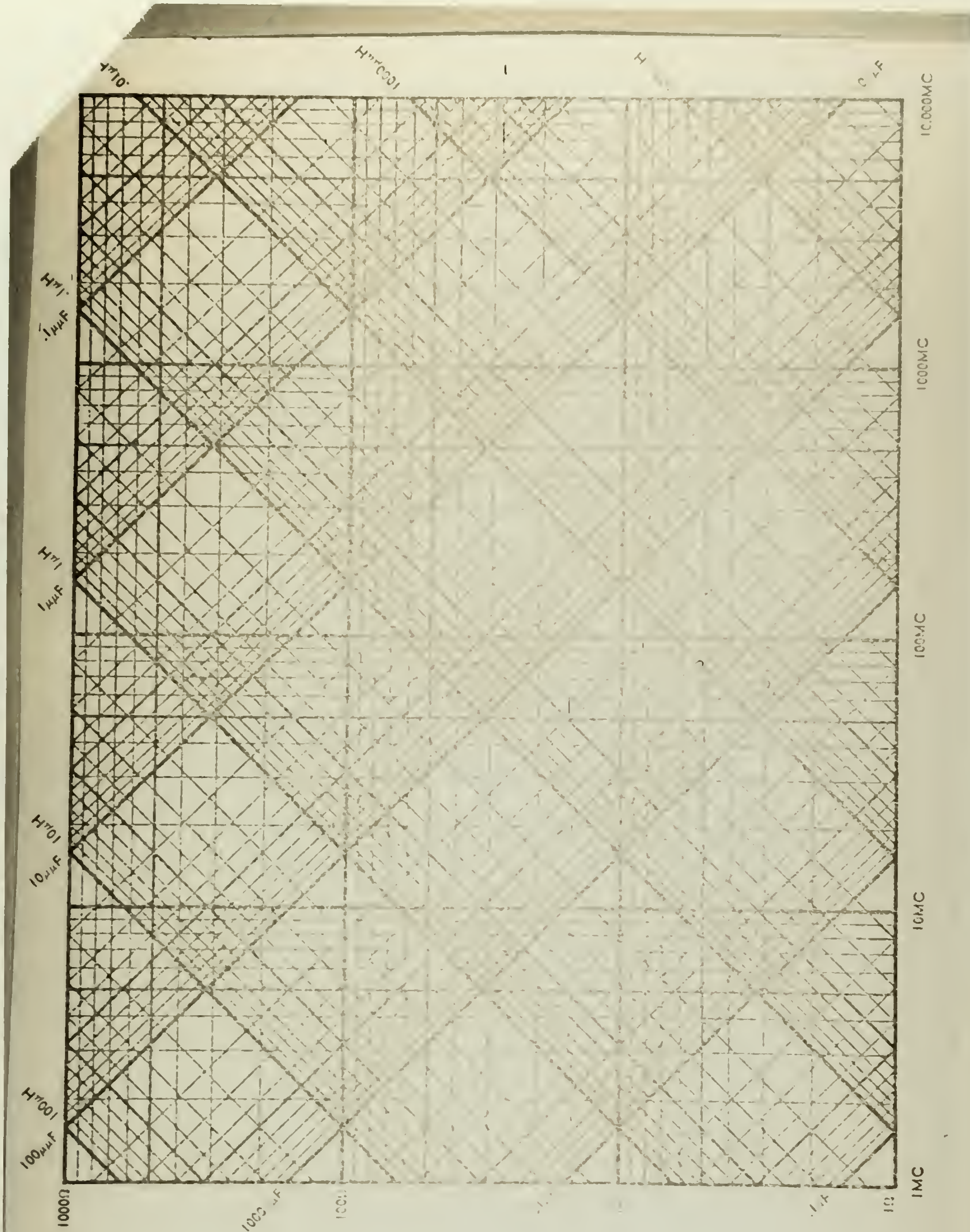


Figure 6

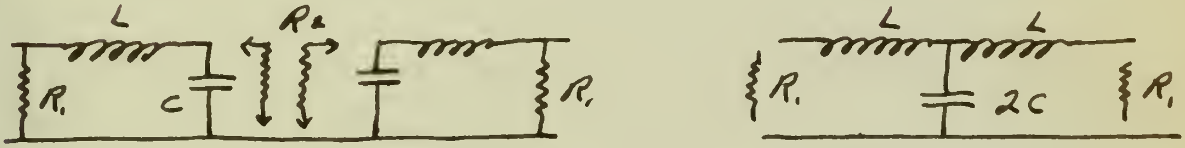
(11) $f_r = \frac{1}{2\pi\sqrt{LC}}\sqrt{\frac{R_L^2 - L/C}{R_C^2 - L/C}} = \frac{1}{2\pi\sqrt{LC}}\sqrt{\frac{R_L^2 - Z_0^2}{R_C^2 - Z_0^2}}$

(12) $f \approx \frac{1}{2\pi\sqrt{LC}}$

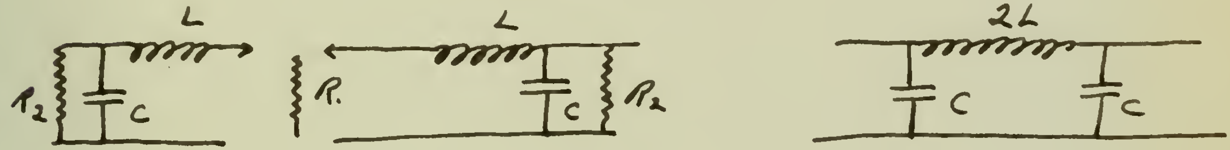
(13) $Z_0 = \sqrt{L/C} = \sqrt{2}\,R_1$

(14) $f_r = \frac{1}{2\pi\sqrt{LC}}\sqrt{\frac{R_1^2-(\sqrt{2}R_1)^2}{2R_1^2-(\sqrt{2}R_1)^2}} = \frac{1}{2\pi\sqrt{LC}}\sqrt{\frac{R_1^2-2R_1^2}{4R_1^2-2R_1^2}} = \frac{.707}{2\pi\sqrt{LC}}$

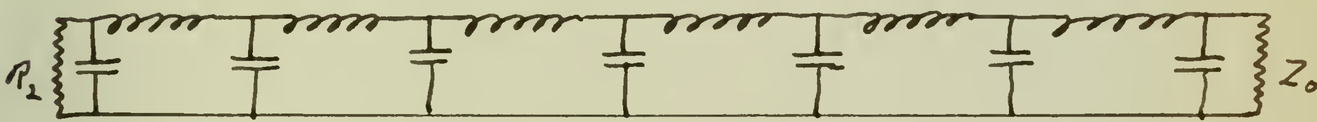
(15) $f_x = \sqrt{1+\cot^2\phi}\; f_r = \sqrt{1^2+\frac{1}{\tan^2\phi}}\; f_r = \sqrt{1+\left(\frac{1}{2Q}\right)^2}\; f_r$



equivalent T



equivalent π



artificial line or ladder equivalent

Figure 4

artificial line will reflect and form standing waves in exactly the way they are formed on an actual transmission line. In some of the circuit examples given later the line must be kept flat, in others the line must have standing waves to function.

Figure (4) shows a resistive load at both ends of the section. It should be kept in mind that one of these is the load, the other the driving point impedance. Only one represents a load, the choice of which is up to the designer.

Appendix A-I gives examples of the use of the chart (Figure (3)). Figure (5) is a part of this appendix. Figure (6) is given here as an auxiliary chart to figure (3).

CHAPTER II

GRAPHICAL SOLUTIONS

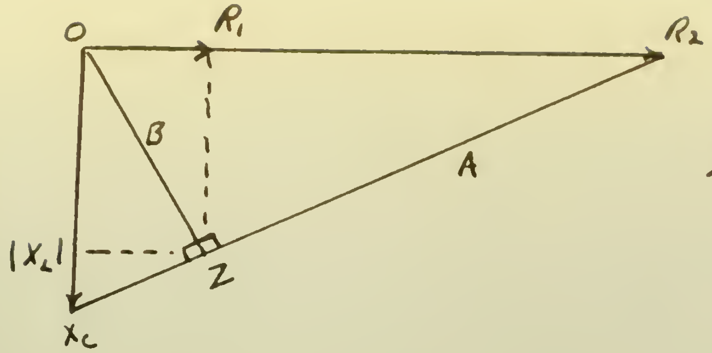
Several graphical solutions to the problem are possible. These are found in a variety of references, chiefly the articles by Paine,[12] Bruene[13][15] and Storch.[14] The use of the Smith and Bode charts is somewhat original.

The basis of all geometric solutions is to convert parallel to series equivalent circuits by means of similar triangles, and then add series elements vectorially. In figure (7), a parallel RC circuit is assumed, and the reactances plotted out. The reactances are connected by the construction impedance line A and a perpendicular is dropped to it from the origin. The calculations with figure (7) show the line B to be the magnitude and phase of the equivalent series combination.

If the equivalent values are to be used in calculations and not necessarily as a part of the plot necessary for further graphical solution, the solution can be made on a Smith circle chart with normalized values. Use of this chart for this purpose will be taken up later.

Another graphical solution which may be needed, particularly in obtaining transfer functions, is multiplication and division of complex numbers. Appendix A-II shows a graphical method of accomplishing this which may simplify calculations where a large number of values must be tried.

Referring to figure (9(a)), the procedure for setting up the graphical solution is as follows: Lay out on the horizontal axis the value of R_2, and on the negative vertical axis, the value of X_c. If then,



$$\bar{Z} = \frac{\bar{R}_c \bar{X}_c}{\sqrt{\bar{R}_c^{\,2} + \bar{X}_c^{\,2}}}$$

$$\frac{OZ}{OR_2} = \frac{OX_c}{X_c R_2}$$

but $X_c R_2 = \sqrt{(\overline{OR_2})^2 + (\overline{OX_c})^2}$

hence

$$\frac{\sqrt{(\overline{OR_2})^2 + (\overline{OX_c})^2}}{OR_2} = \frac{OX_c}{OZ_T}$$

and

$$\overline{OZ} = \frac{(\overline{OX_c})(\overline{OR_2})}{\sqrt{(\overline{OR_2})^2 + (\overline{OX_c})^2}} = \frac{\bar{X}_c \bar{R}_2}{\bar{R}_2 + \bar{X}_c}$$

Figure 7

a circle with diameter equal to R_2 is drawn the intersection of this circle with the line A is the point of the Z vector equivalent to the parallel combination of X_c and R_2. A perpendicular to this will give the equivalent values of X_{cs} and R_1 as shown. To make a perfect match at one frequency, X_L is chosen equal to X_{cs}. Note that the vector sum of X_L and B then becomes the vector R_1 which has zero power factor.

When X_L and R_1 are known, they are plotted as shown in figure (9(b)). Since the line C must be perpendicular to line B', the value of X_{Lp} and R_2 can be determined by the intersections of this line with the horizontal and vertical axes, X_c is then made equal to X_{Lp} and again the vector sum of B' and X_c is the pure resistance R_1.

To graphically solve a pi network, draw two circles of diameter R_2 and R_2' as shown in figure (10). The value of R_1 is then selected by means of equation (10) from the desired Q. Since R_1 is common, a perpendicular from it to the circles will determine two values of X_L, and the two lines A and A' will determine the values of input and output capacity. Referring to the figures (9) and (10), the Q of the circuit is represented by the slope of the lines A and C. This being apparent from equation (9). The Q of the pi circuit is modified by the presence of two R_1 values, two X_L values and two X_{cs} values. The overall Q is determined from the sum of these values again plotted and the slope taken as Q, or calculated from equation (9).

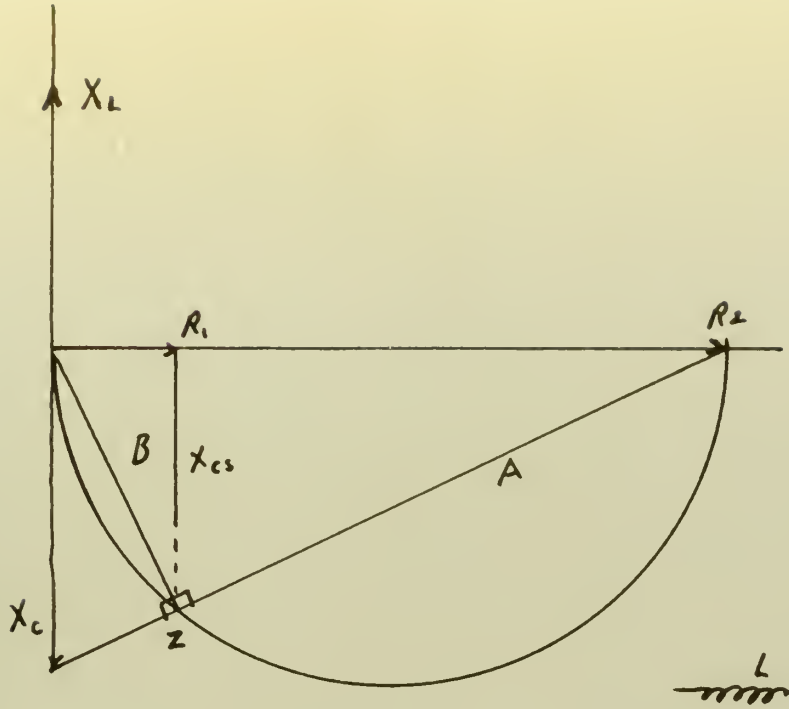
Figure (11(a)) shows a graphical plot of the T match. R_1 and R_1' may be different, but R_2 is a common element. Note that the B and B' vectors lie on a common circle. From this plot two values of X_c are

a circle with diameter equal to R_g is drawn, the intersection of this circle with the line A is the point of the vector representing the parallel combination of R_g and L_g. A line perpendicular to this will give the [illegible] line of Y_{oo} [illegible] figure plane. [illegible] at [illegible] frequency, Y_i is common [illegible] Y_o. Note that the vector [illegible] of Y_o [illegible] then becomes the vector Y_1 which [illegible] figure.

When Y_i and Y_1 are known, they are plotted as shown in figure (9(b)). Since the line C must be perpendicular to line A', the value of R_g and L_g can be determined by the intersection of this line with the horizontal and vertical axes, Y_o is then [illegible] equal to [illegible] and makes the vector sum of A' and Y_o is the [illegible] vector Y_i.

To graphically solve [illegible], [illegible] the circles of diameter R_g and [illegible] as shown in figure (10). The value of Y_1 is then obtained by means of equation (11) from the [illegible]. Since Y_1 is common, a perpendicular from it to the circles will determine two values of Y_1, and the two lines A and A' will determine the values of input and output capacity. Referring to the figures (9) and (10), the [illegible] of the circuit is represented by the slope of the lines A and C. This [illegible] from equation ( ). The [illegible] of the [illegible] circuit is modified by the presence of two Y_i values, the Y_1 values and the Y_{oo} values. The overall [illegible] is determined from the sum of these [illegible] plotted on the [illegible], or calculated from equation ( ).

Figure (11(a)) shows a graphical plot of the Y vectors. Y_1 and Y_1' may be different, but [illegible] is a common element. Note that the A and A' vectors lie on a common circle. From this plot the values of Y_o are



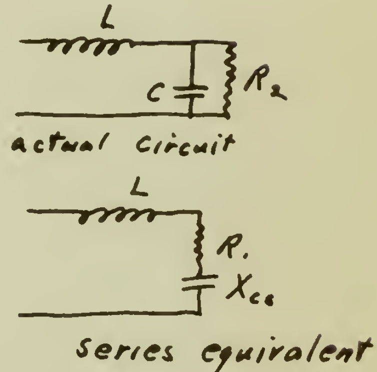
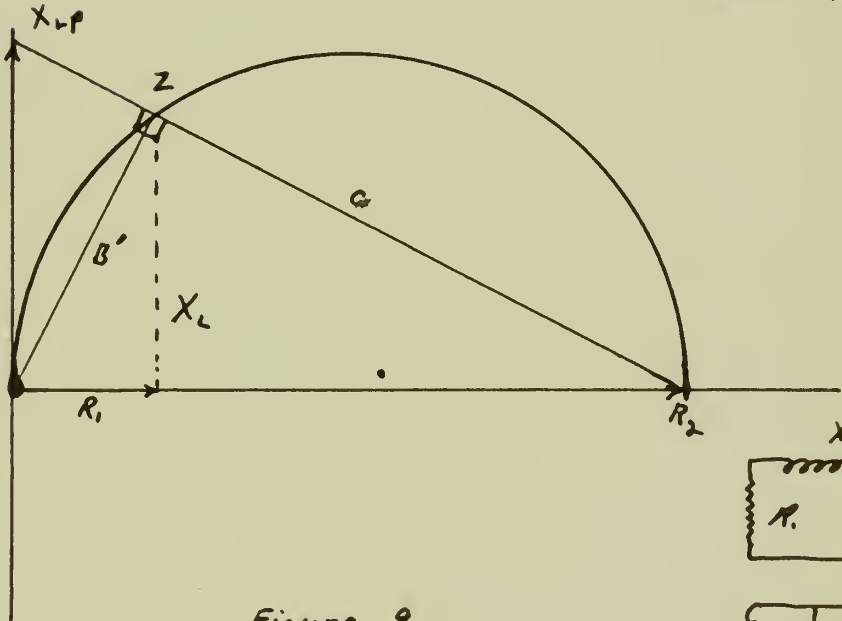


Figure 8



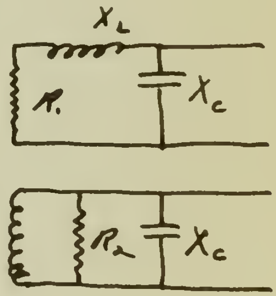
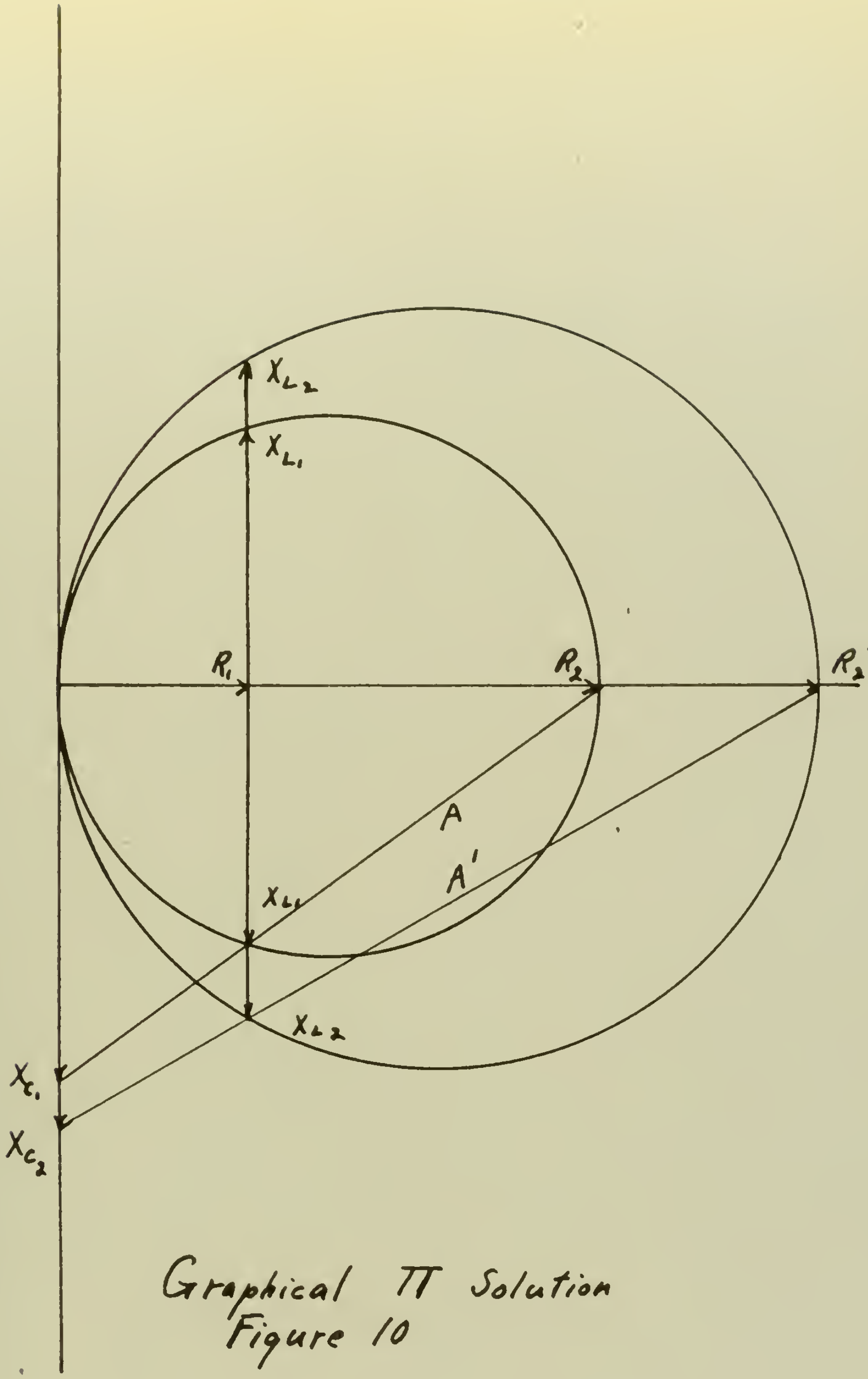


Figure 9

found which may be paralleled graphically as in figure (12) to give the equivalent parallel reactance. If this is done to X_L, X_c and R_1 separately, then the whole plotted as a single LCR combination, the overall Q of the T match can be determined.

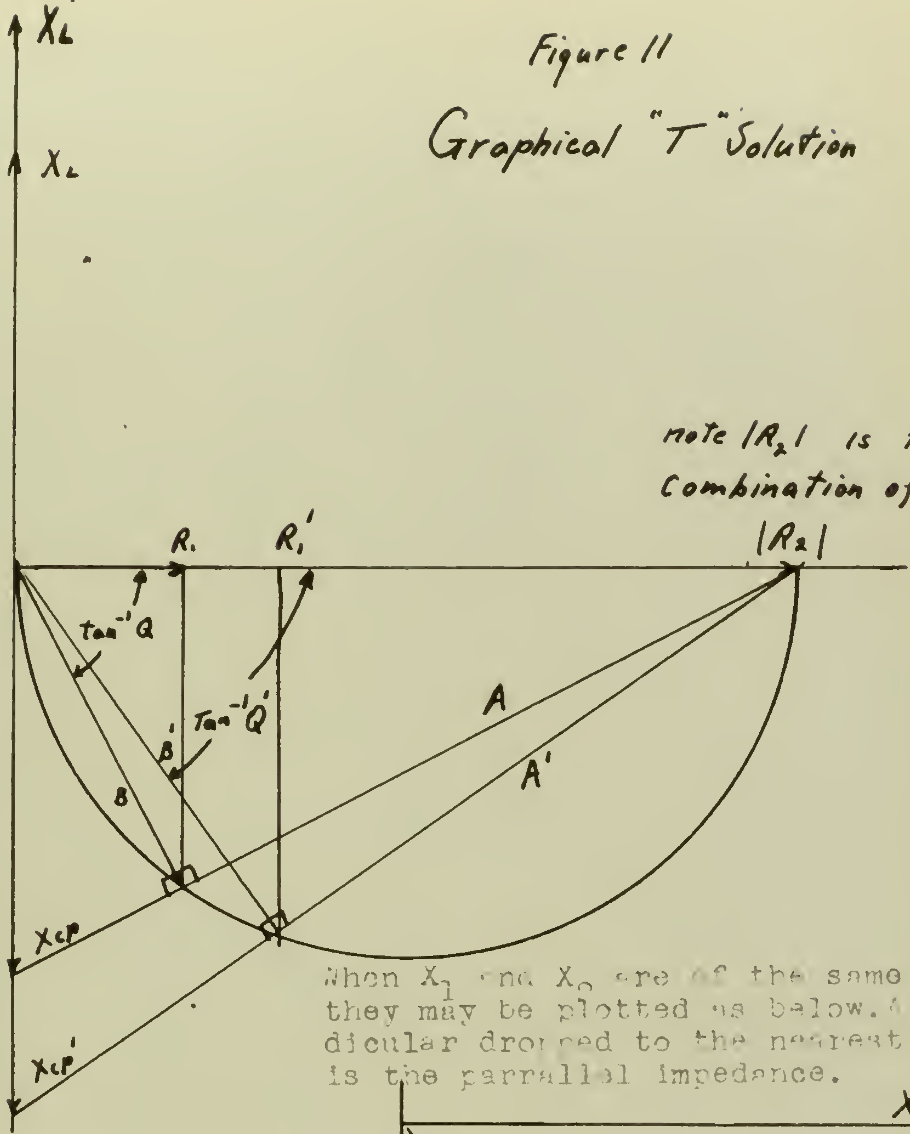
All three circuits, L, T and pi, when plotted out and combined, yield the same basic LCR circuit which leads one to expect all three will have identical transfer functions. This is not a valid assumption, however, as will be shown in the next chapter.



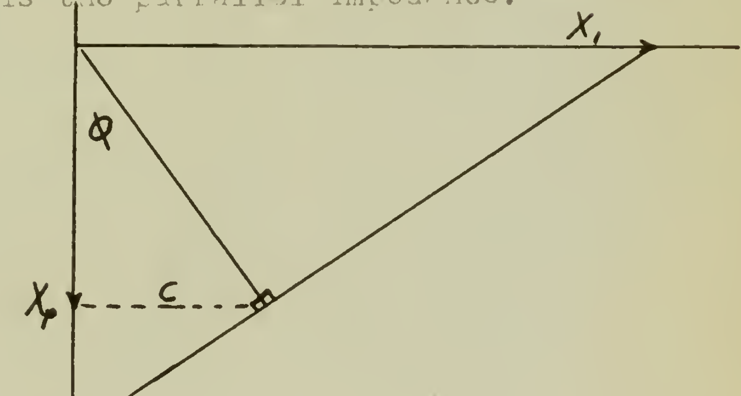
Graphical π Solution
Figure 10

Figure 11

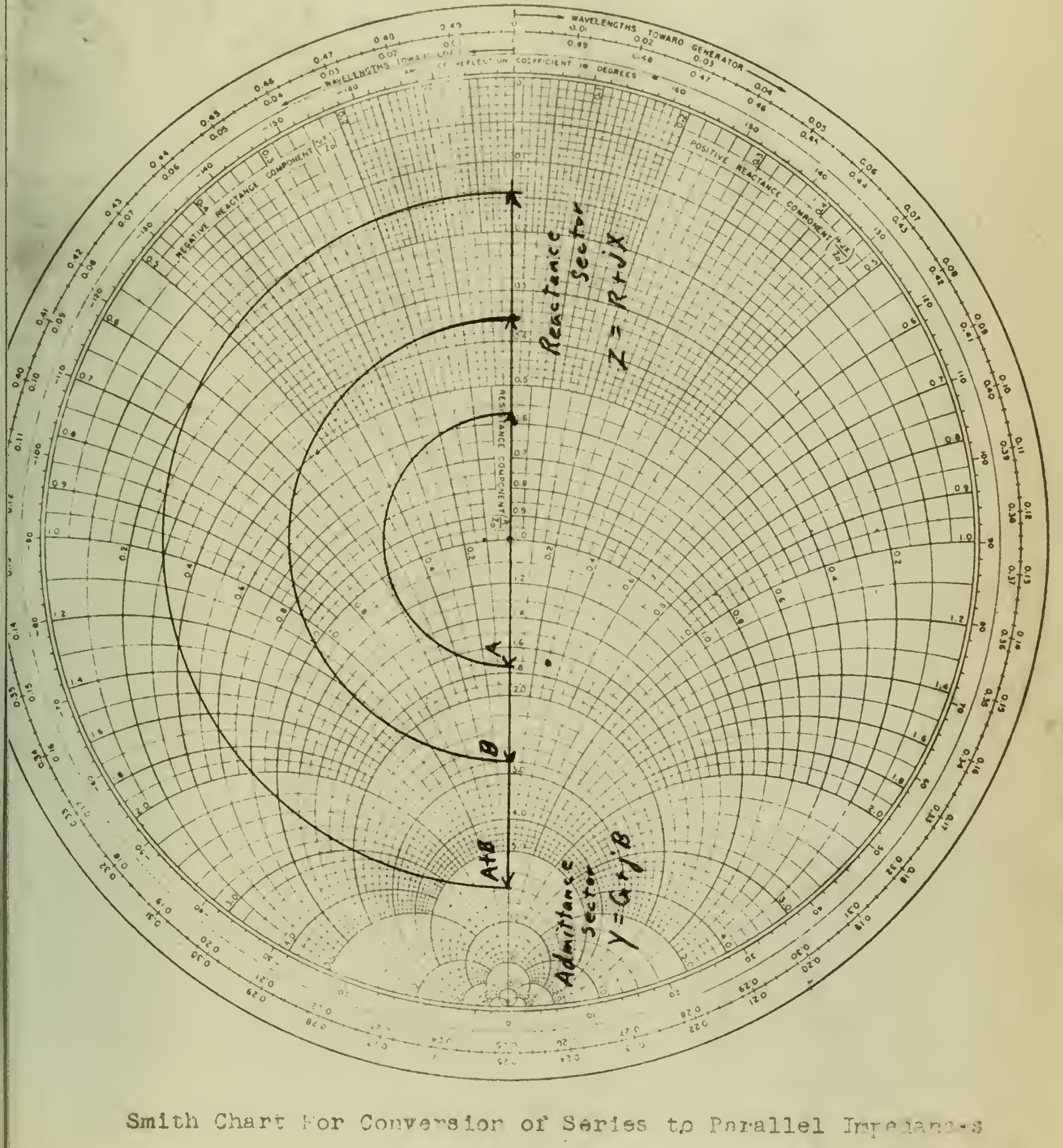
Graphical "T" Solution



When X_1 and X_2 are of the same phase they may be plotted as below. A perpindicular dropped to the nearest side is the parrallel impedance.



Normallize the values of the impedances to 1. Locate the complex impedance on the chart, draw a line thru the center an equal distance out- this locates the complex admittance. Add complex admittances to get the parallel equivalent. No particular sector is for admittance or impedance as shown here-- impedances usedin this example are less than 1. Once complex sum is obtained , plot back thru center to get complex impedance.



Smith Chart For Conversion of Series to Parallel Impedances

Figure 12

CHAPTER III

TRANSFER FUNCTIONS OF L, T AND PI NETWORKS

Figure (13) shows the L match. The transfer function is developed as shown to yield equation (16) and a G(s) of the form of equation (16a). The term, (s), is used to denote Laplace transforms are being used. Most writers on networks employ the term, (p), which has the same meaning. G(s) is a second order function which will plot in the left hand(s) plane denoting a damped oscillatory circuit. The input impedance has poles at zero and infinity with an intermediate zero. The transfer function has an intermediate pole and a zero at infinity. Whether or not a zero is assumed at zero is a matter of interpretation. The poles and zeros plot as in figure (13).

Figure (14) illustrates the pi match and shows the derivation of its transfer function. The transfer function here is third order denoting a maximum of 270 degrees phase shift and an additional zero in the input impedance.

Figure (15) illustrates the T match and shows the derivation of its transfer function. The transfer function is similar to that of the pi match except that it is multiplied by (s) which indicates a 90 degree phase rotation. Accordingly the poles and zeros plot as shown and appear shifted along the axis.

It will be noted the plots of input Z very closely resemble those of transmission lines of various lengths allowing for a sharp cutoff frequency which limits the number of poles and zeros to those shown.

[illegible]

[illegible]

Figure (11) [illegible] as shown in this [illegible] form of equation (iii). The term [illegible] (u), is used to [illegible] [illegible] which has the same [illegible].

(u) is a second order function [illegible] will plot in the left half (u) plane denoting a damped oscillatory circuit. The input impedance has poles at zero and infinity, with an intermediate zero. The transfer function has an intermediate pole and a zero at infinity. Whether or not a zero is obtained [illegible] is a matter of [illegible]. The poles and zeros plot as in Figure (12).

Figure (13) illustrates [illegible] which shows the derivation of the transfer function. [illegible] zero in the input impedance.

Figure (14) illustrates [illegible] and shows the derivation of the transfer function. The transfer function is similar to that of [illegible] except that [illegible] phase rotation. Accordingly the poles and zeros plot as shown [illegible] shifted from the [illegible].

It will be noted that [illegible] very closely, [illegible] of transmission lines of various [illegible] not a sharp cutoff frequency which limits the number of poles and zeros to those shown.

Figures (16), (17), (18), (19), and (20) show the three basic sections plotted for an arbitrary Q. In order to condense the curves, logarithmic scales are used.

Calculating the amplitude and phase variations of the transfer functions is a long laborious process. To condense this time and permit experiments where parameters are varied readily, use was made of the Analog computer. The computer available at the United States Naval Postgraduate School is a 10 amplifier unit built by Boeing Aircraft Corporation. This unit multiplies and integrates, but is unstable when differentiating. For this reason the units are arranged as shown so that no differentiation is required.

The output is recorded on a two channel brush recorder with a maximum reliable response of about 60 cycles. For this reason the frequency is scaled with a resonance at about 15 cycles. This makes it possible to use readily available R and C values in the computer, and yields a form of universal curve. The values used are $C = 10^{-3}$, $L = .1$ and R according to Q. X_C at 15 cycles is 10 ohms.

The set up for the analog computer used is given in Appendix A-III.

Several conclusions may be drawn from the transfer functions and curves:

1. When sections are cascaded, the resulting transfer function is not $[G(s)]^n$. An artificial transmission line results with standing waves.

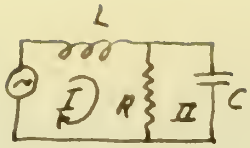
2. A special case exists where Q = 0, the transfer function is $[G(s)]^n$.

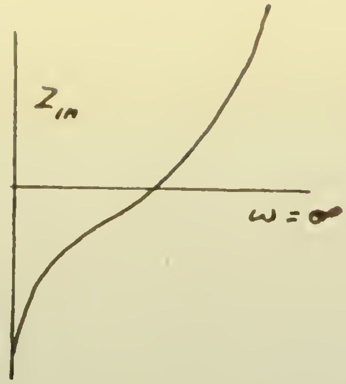
[illegible]
sections [illegible]
logarithmic scales are used.
[illegible]
functions [illegible]
[illegible] of the analog computer. [illegible]
[illegible]
eight [illegible]
and differentiating. [illegible]
so that no differentiation is involved.
The output is recorded [illegible]
maximum reliable response [illegible] cycles. [illegible]
frequency is [illegible] cycles. [illegible]
possible to [illegible]
yields [illegible]
L = .1 henry [illegible]
[illegible]
general conclusions [illegible]

CURVES

1. [illegible]
and [illegible]
straight [illegible]

2. [illegible]
is [illegible]





Loop Equations

$$(L_1 s + R_2) I_1(s) - R_2 I_2(s) = E_1(s)$$

$$-R_2 I_1(s) + \left(R_2 + \tfrac{1}{Cs}\right) I_2(s) = 0$$

which yields

$$I_2(s) = \frac{E_1(s) R_2 s}{L_1 R_2 s^2 + \left(R_2^2 + \frac{L_1}{C}\right) s + \frac{R_2}{C}}$$

$$E_2(s) = I_2(s) X_c = \frac{I_2(s)}{Cs}$$

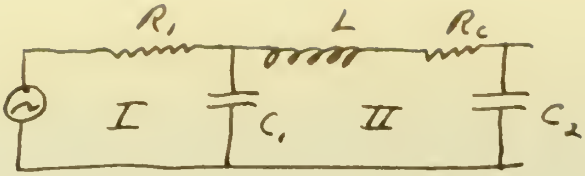
$$\frac{E_2(s)}{E_1(s)} = \frac{R_2}{\left(\frac{L_1}{C} R_2\right) s^2 + \left(\frac{R_2^2}{C} + \frac{L_1}{C^2}\right) s + \frac{R_2^2}{C^2}} \qquad (16)$$

which is of the form

$$G(s) = \frac{a_0}{s^2 + b_1 s + b_0} = \frac{a_0}{(s+\alpha)(s+\beta)} \qquad 16(a)$$

Figure B

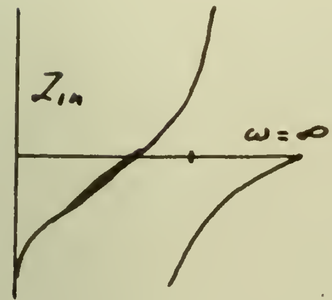
Transfer function of L Section



π network driven by Source with impedance R_1. Q is limited by R_c

Loop Equations

$$\left(R_1 + \frac{1}{C_1 s}\right) I_1(s) - \frac{1}{C_1 s} I_2(s) = E_1(s)$$



$$-\frac{1}{C_1 s} I_1(s) + \left(\frac{1}{C_1 s} + \frac{1}{C_2 s} + L_1 s + R_c\right) I_2(s) = 0$$

$$I_1(s) = \frac{\left(\frac{1}{C_1 s} + \frac{1}{C_2 s} + L_1 s + R_c\right) I_2(s)}{1/C_1 s}$$

$$= 1 + \frac{C_1}{C_2} + L_1 C_1 s^2 + R_c C_1 s$$

$$\left(R_1 + \frac{1}{C_1 s}\right)\left(1 + \frac{C_1}{C_2} + L_1 C_1 s^2 + R_c C_1 s\right) I_2(s) - \frac{1}{C_1 s} I_2(s) = E_1(s)$$

$$R_1 + \frac{R_1 C_1}{C_2} + R_1 L_1 C_1 s^2 + R_2 R_1 C_1 s + \frac{1}{C_2 s} + L_1 s + R_2 = \frac{E_1(s)}{I_2(s)}$$

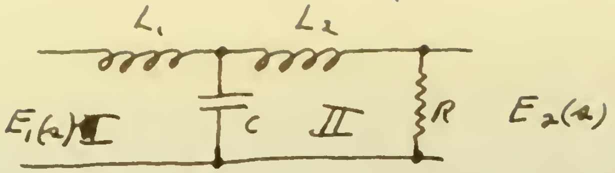
$$I_2(s) = \frac{E_2(s)}{X_c}$$

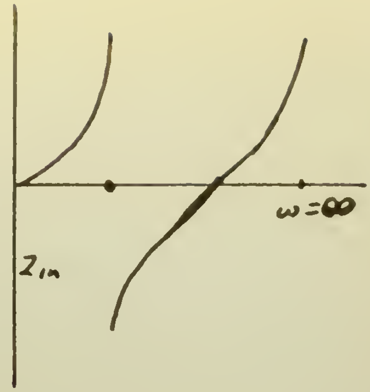
$$\frac{E_2(s)}{E_1(s)} = \frac{1}{R_1 L_1 C_1 C_2 s^3 + (R_1 R_2 C_1 C_2 + L_1 C_2) s^2 + (R_1 C_2 + R_1 C_1 + R_2 C_2) s +} \qquad (17)$$

which is of the form

$$G(s) = \frac{a_0}{s^3 + b_2 s^2 + b_1 s + b_0} \qquad (17a)$$

Figure 14- Transfer function of π Section.





Loop Equations

$$\left(L_1 s + \frac{1}{Cs}\right) I_1(s) - I_2(s)\left(\frac{1}{Cs}\right) = E_1(s)$$

$$-\left(\frac{1}{Cs}\right) I_1(s) + \left(L_2 s + R + \frac{1}{Cs}\right) I_2(s) = 0$$

$$I_1(s) = I_2(s)\left[Cs\left(Ls + R + \frac{1}{Cs}\right)\right]$$

$$E_1(s) = I_2(s)\left[Cs\left(Ls + R + \frac{1}{Cs}\right) + Ls + \frac{1}{Cs}\right]$$

$$E_2(s) = I_2(s) R$$

$$\frac{E_2(s)}{E_1(s)} = \frac{R}{\left[Cs\left(Ls + R + \frac{1}{Cs}\right) + Ls + \frac{1}{Cs}\right]}$$

$$= \frac{Rs}{LCs^3 + (RC + L)s^2 + s + \frac{1}{C}} \qquad (18)$$

which is of the form

$$G(s) = \frac{a_1 s}{s^3 + b_2 s^2 + b_1 s + b_0} \qquad (18(a))$$

Figure 15 - Calculation of G(s) for T Match.

Figure 16

Transfer function of "L" Network
Arbitrary Q

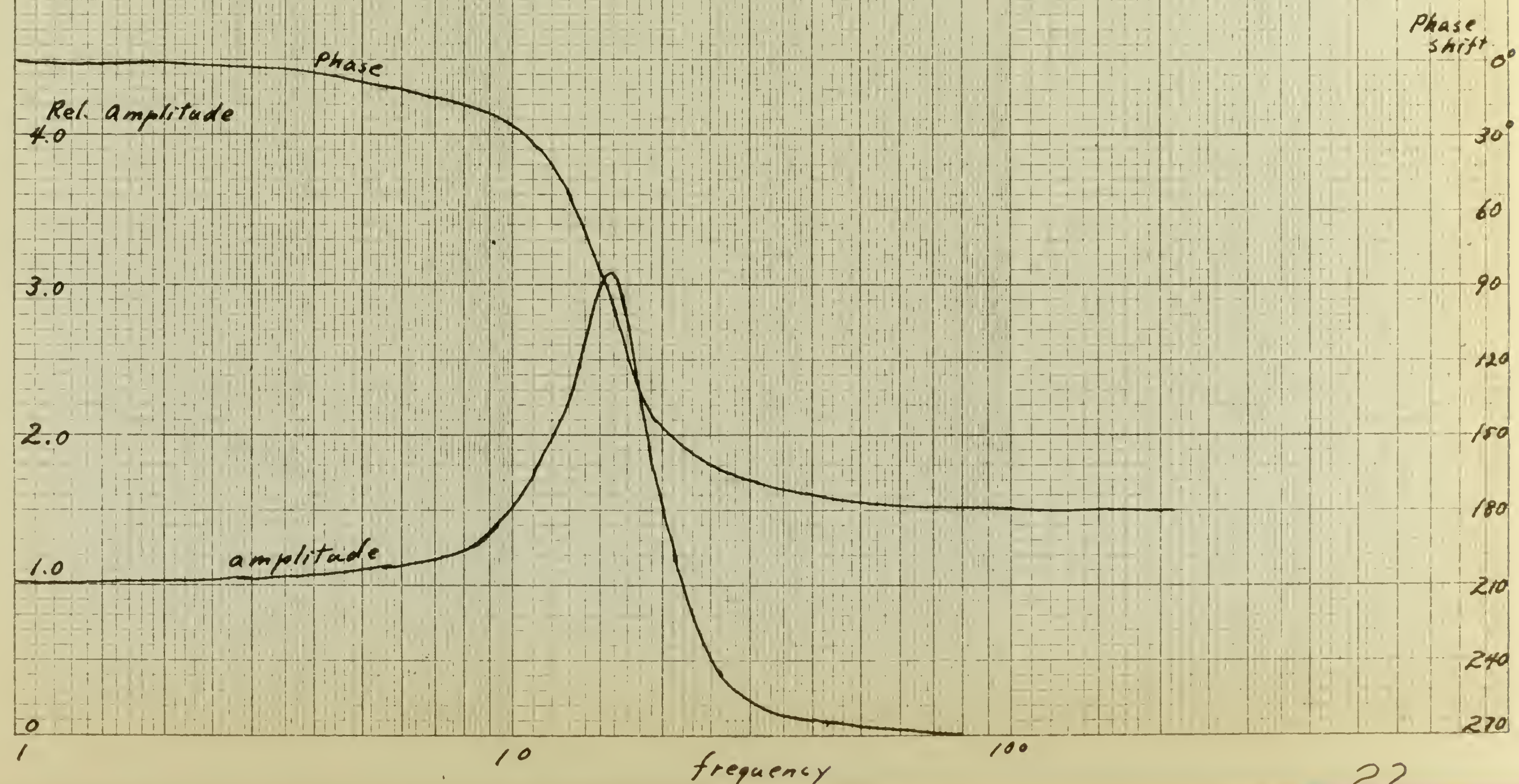


Figure 17

Transfer function of T network
Arbitrary Q

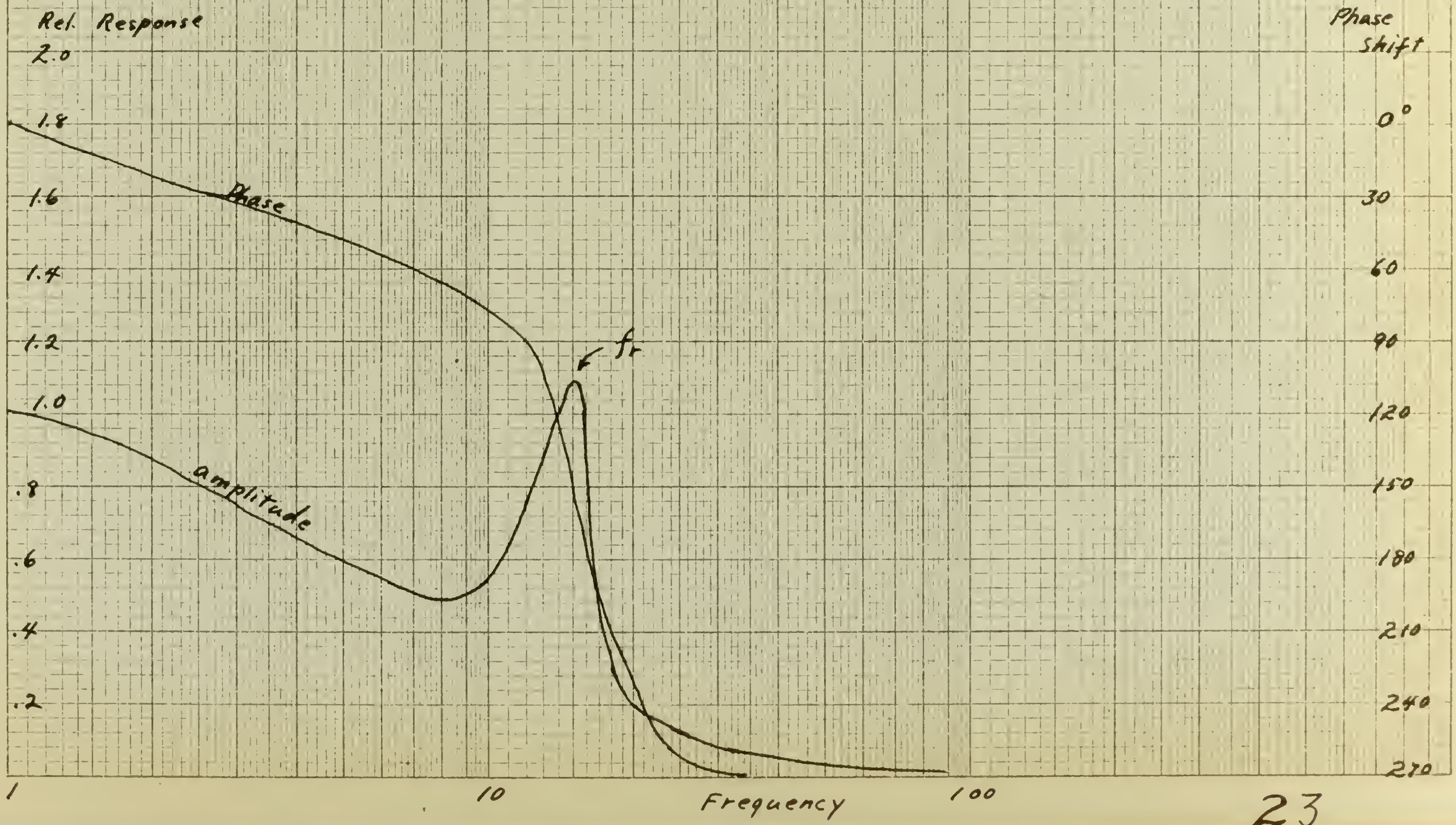
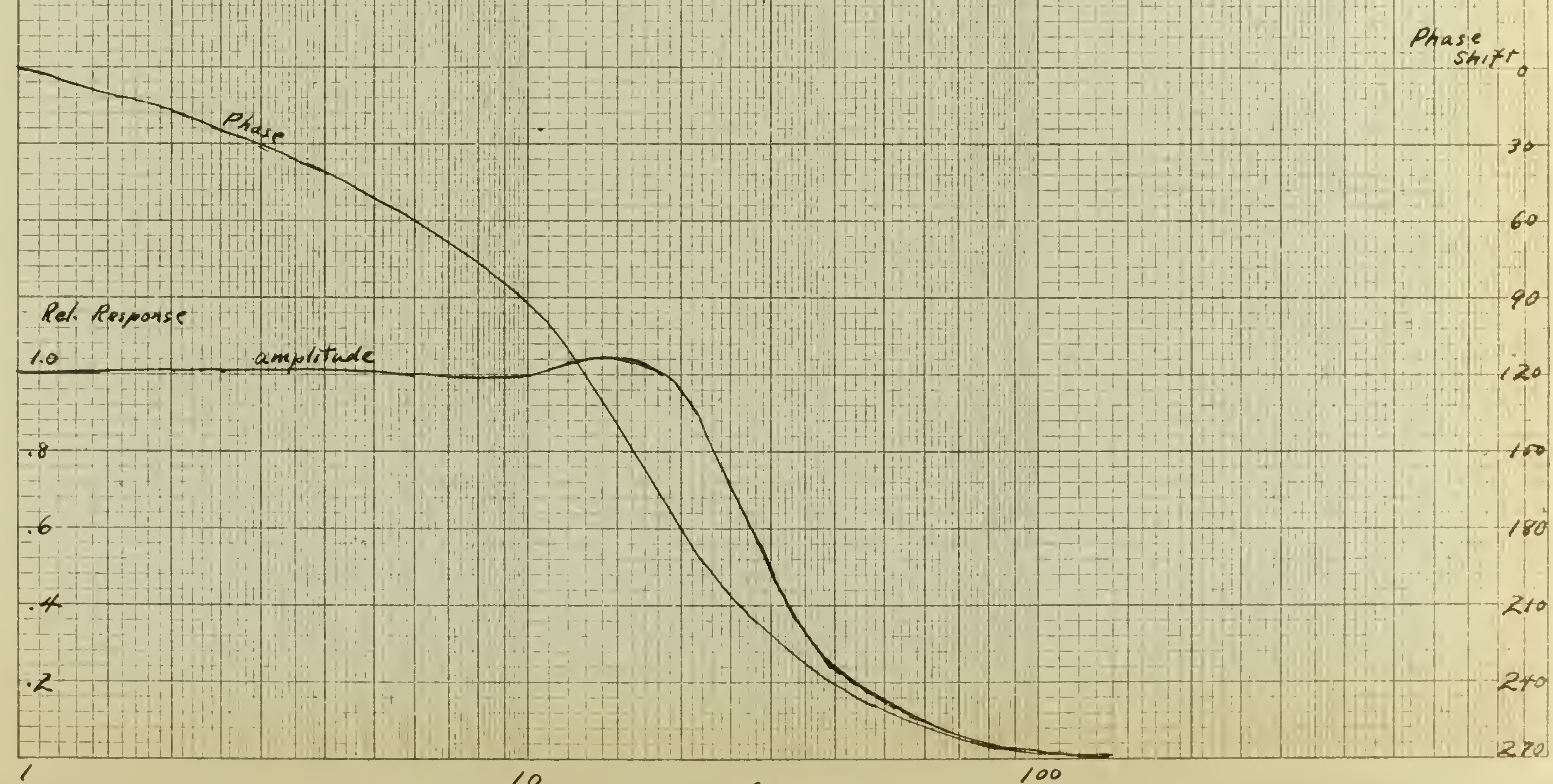


Figure 18

Transfer function of π network

Q arbitrarily set for optimum Video Compensation



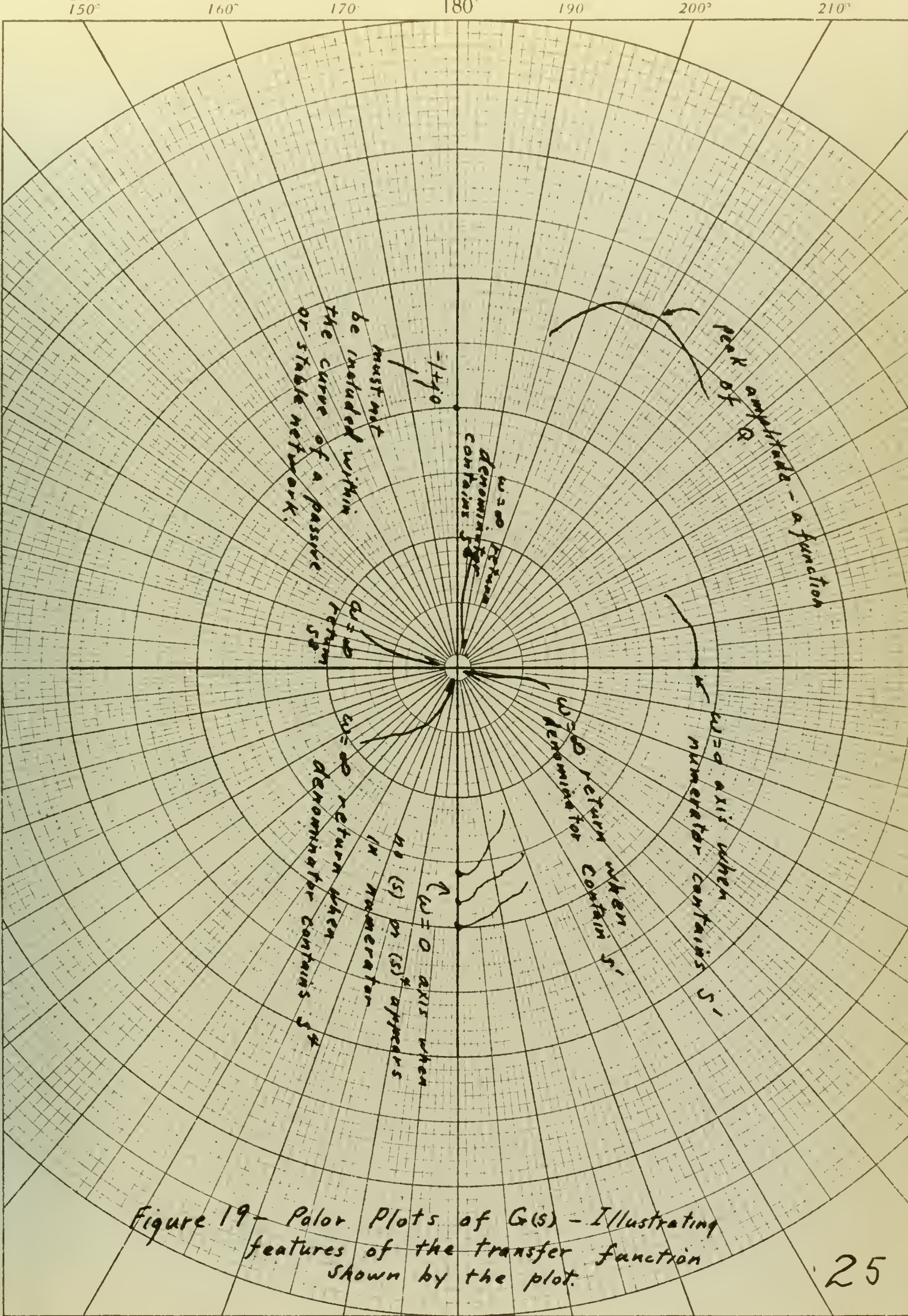


Figure 19 - Polar Plots of G(s) - Illustrating features of the transfer function shown by the plot.

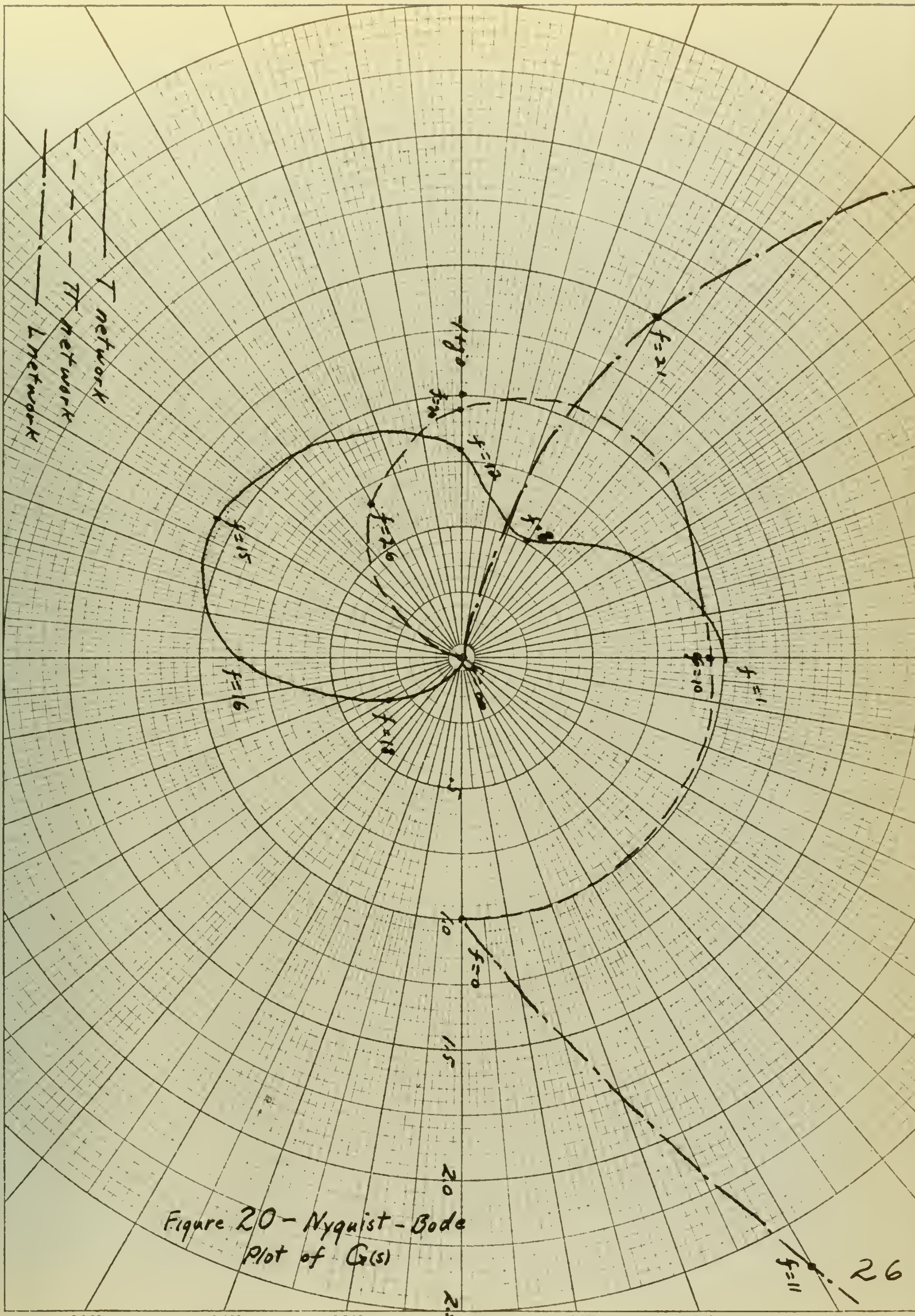


Figure 20 – Nyquist - Bode Plot of G(s)

3. The transfer functions of symmetrical pi and T networks are bilateral, while that of the L network is not.

4. Voltages at resonance is the pi curcuit divided in inverse proportion to the ratio of input to output capacitance.

5. Voltages divide as the square roots of the input and output impedances.

6. The following equation holds:

$$\rho = \frac{E_{max}}{E_{min}} = \sqrt{Q^2+1} = |G(s)|_{max}$$

7. Universal curves are possible for the L network displaying amplitude and phase for various Q's.

8. Universal curves for the pi and T networks are impractical because of the many variables which alter them.

9. Where Q exceeds zero, the impedance match obtained by pi, T and L networks is reliable only at resonance.

CHAPTER IV

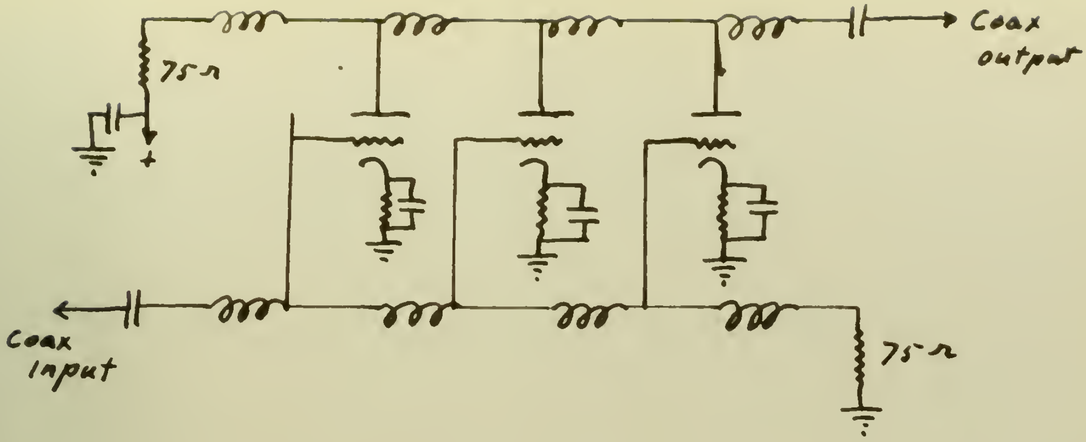
DISTRIBUTED AND CASCADED AMPLIFIERS

One type of broadband amplifier patented by Percival[56] in England and developed in this country by Spencer, Hewlett and Kennedy is the distributed amplifier which makes use of artificial transmission line sections to reduce the effect of tube capacity, provide a constant load impedance to the tubes, and a relatively constant gain over a broadband of frequencies up to about 220 mc. Using special tubes, Melpar Incorporated has developed an amplifier of this type usable to about 1,000 mc.

The limit in gain for amplifiers of this type is approximately 20 db, and while they are used in special applications such as millimicrosecond oscillography, for television they have largely been superseded by a series of cascaded triodes with either pi or M derived coupling sections. Triode boosters for television using 4 dual triodes give a 24 db gain with a noise figure of 12 db over the television bands. Pentode distributed amplifiers use 8 pentodes give 22 db of gain, with a 20 db noise figure over the same band.

Figure (21) shows the basic Spencer-Kennedy[48][49] circuit. In designing this circuit, the L is usually made a continuous coil on a long rod. The tube elements furnish the capacity - except that some slight additional capacity is needed in the plate circuits to make the grid and plate artificial lines identical.

There are more involved concepts to explain the circuit, but the simplest is that the tubes are all in parallel, and the only function



For Pentodes

$$G_m' = n \times G_m$$

$$R_L = \tfrac{1}{2} Z_0 = 37.5\,\Omega$$

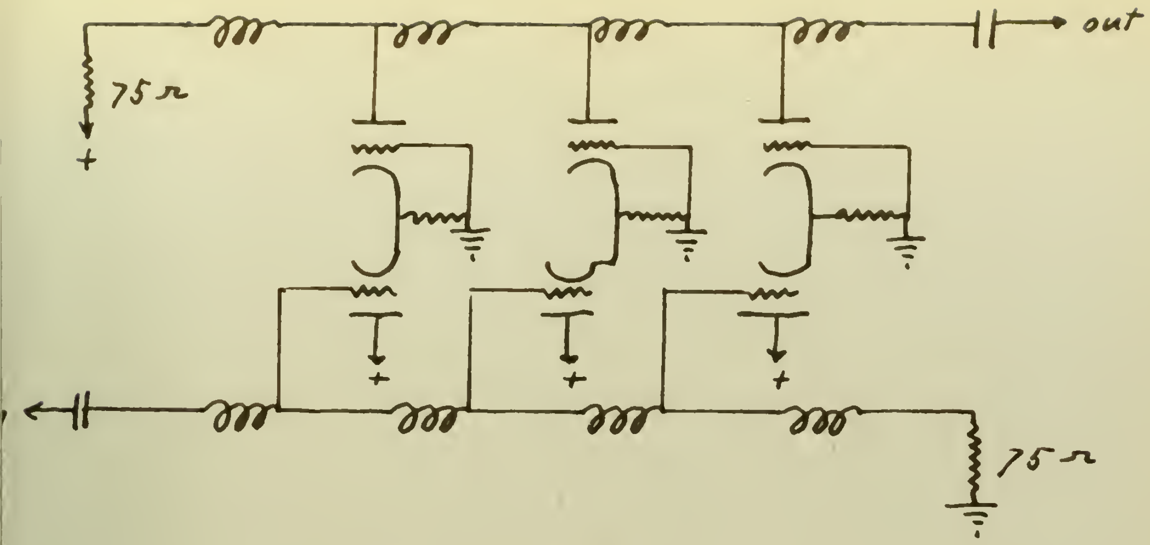
Assume 8 - 6AK5's

$$G_m = 5{,}000\ \mu\text{mhos}$$

$$G_m' = 40{,}000\ \mu\text{mhos}$$

$$\text{Gain} = R_L' G_m' = 37.5 \times 40 \times 10^{-3} = 15 = 23\,\text{db}$$

Figure 21 - The Distributed Amplifier



$$\text{Gain} = \frac{K \mu R_L'}{R_p' + R_L'}$$

K = a loss factor due to paraphase coupling $\approx .7$

Assume $\mu = 40$

$R_p = 5{,}000\ \Omega$

8 Tubes

$R_p' = 625$

$R_L = 37.5$

$$G = \frac{40 \times 37.5 \times .7}{625 + 37.5} = 1.6 = 4\text{db}$$

Figure 22 - The Paraphase Amplifier

of the line is to isolate tube capacities and give a constant load over the pass band.

Because of the high equivalent noise resistance of the pentodes, (about four times that of a VHF triode), triodes are preferable, but the great difficulty in their use is the need for neutralization. Melpar has devised a circuit using special pencil triodes which is usable to 1000 mc, and does not require neutralization. This circuit is a cathode follower driving a grounded grid section. Melpar prefers to call it a paraphase circuit.

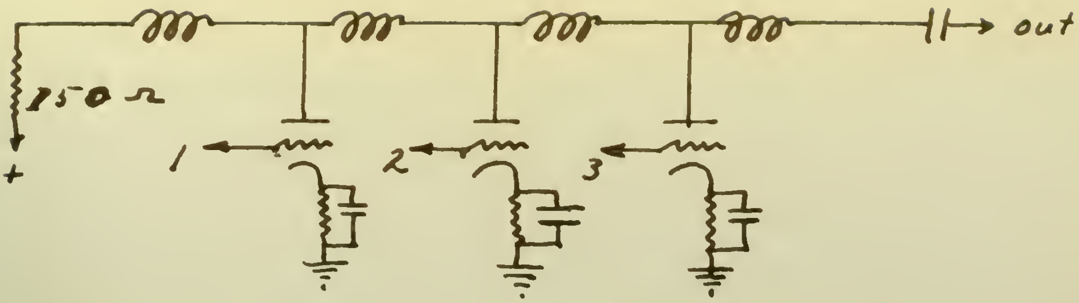
Figure (22) shows the all triode circuit.

Tubes of the 6J6 type are usable to 200 mc. For circuits up to 1000 mc., pencil triodes with a special isolated cathodes are used. In all cases, the filaments must be floating at R. F. by effective R. F. chokes. For the 1000 mc. unit described at the 1953 IRE convention, the inductances were merely straps connecting the tube sockets.

The gain of such a circuit is given in figure (22). It can easily be seen that special tubes with low R_p and high μ are required if appreciable gain is to be realized.

Because of the economy factor in designing an item for home use, as well as for reasons of better performance, present day boosters use pi couplings or M dervied sections between cascaded triode sections.

Distributed amplifiers are not limited to amplifying. They are also useful for mixing and distributing signals. Figure (23) shows a three channel mixer with separate grid inputs, but with all plates forming elements of the line. Pi or L couplers may be used at the grids



For Triodes

$$\text{Gain} = \frac{\mu R_L}{R_L + R_p} = \frac{40 \times 75}{75 + 5{,}000} = .6$$

For Pentodes

$$\text{Gain} = G_m R_L = 5{,}000 \times 10^{-6} \times 75 = .375$$

Figure 23 - The Distributed Mixer

for impedance matching and voltage step up. In this case the Q of the input match is limited by the bandpass required of the input, while the output line usually is made a very low Q or flat.

There is some loss in mixing. Triodes are suitable without neutralizing because the Miller effect is negligible when the gain is less than 1. For television circuits using a 6BQ7 or 6BK7, the line to pass all 13 VHF channels has a maximum Z_0 = 150 ohms, thus R_L = 75 ohms. Triode mixers are to be preferred because of gain and noise figure. Usually they are used in pairs to provide a 300 ohm balanced input and output.

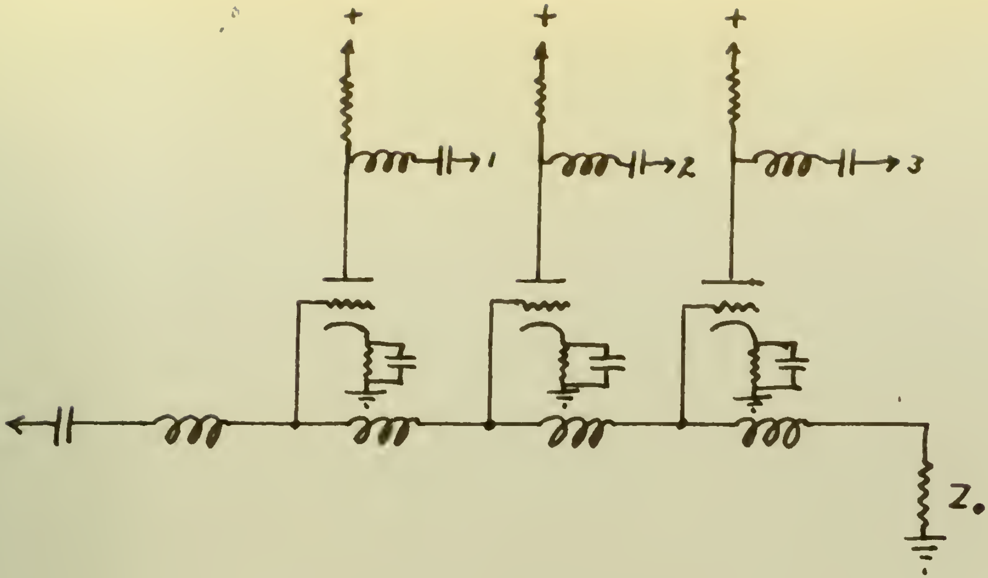
The line may be used in the grid circuit to feed several isolated outlets as in figure (24). Here an artificial line feeds three separate outlets. If the outlet is to utilize only part of the frequencies carried, the L or pi matched plate circuit may have a Q greater than 0. If it is to be broadband, Q is set at 0 or 1. R is large so that its effect on Q is negligible compared to the reflected load of the output. Pentodes may be used singly, or triodes in pairs. The distribution unit has a slight gain.

Triodes have a greater feedback factor so that they offer less isolation to the outputs. For this reason, pentodes are preferred in commercial units. Using pi couplings with a Q of 1 tuned to about channel 11, the pentode units can drive an effective R_L large enough to give a slight gain over most of the spectrum. Because they are single ended however, matching transformers are necessary to feed 300 ohm lines, and the overall effect is a gain of about 1/1. Again, because the line

[illegible]

[illegible]

isolation to the outside. [illegible] commercial units. [illegible] channel [illegible] give a slight [illegible]



For Triodes $\quad$ $\text{Gain} = \dfrac{\mu R_L}{R_L + R_p} = \dfrac{40 \times 150}{150 + 5{,}000} = 1.16$

For Pentodes

$$\text{Gain} = G_m R_L = 5{,}000 \times 10^{-6} \times 150 = .75$$

Figure 24 - Signal Distribution System

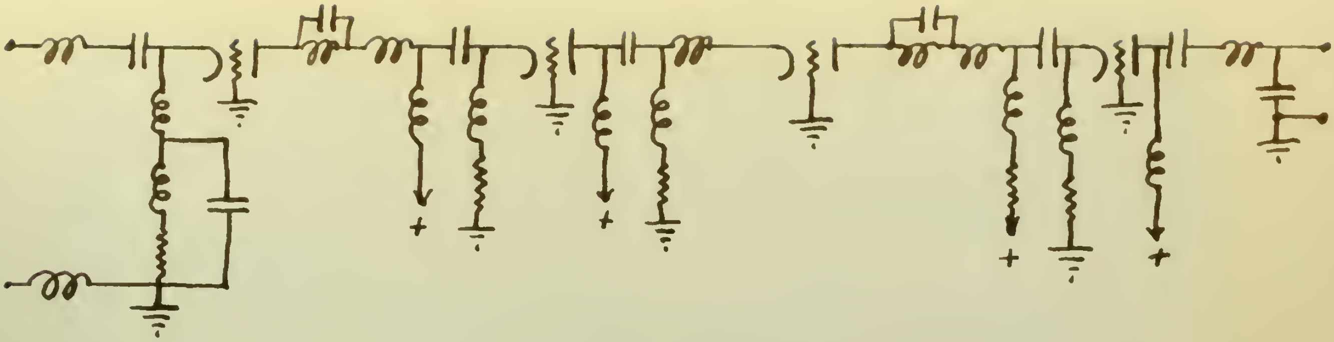


Figure 25 -
Cascaded Triode R.F. Amplifier

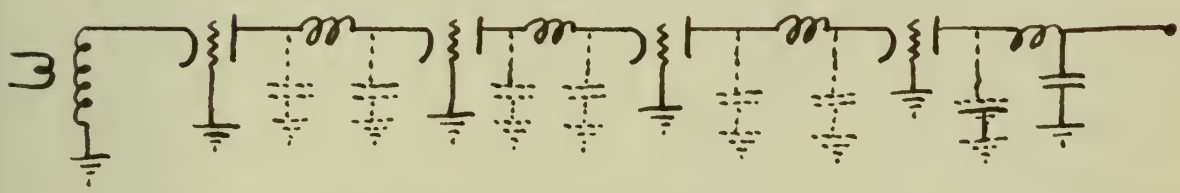
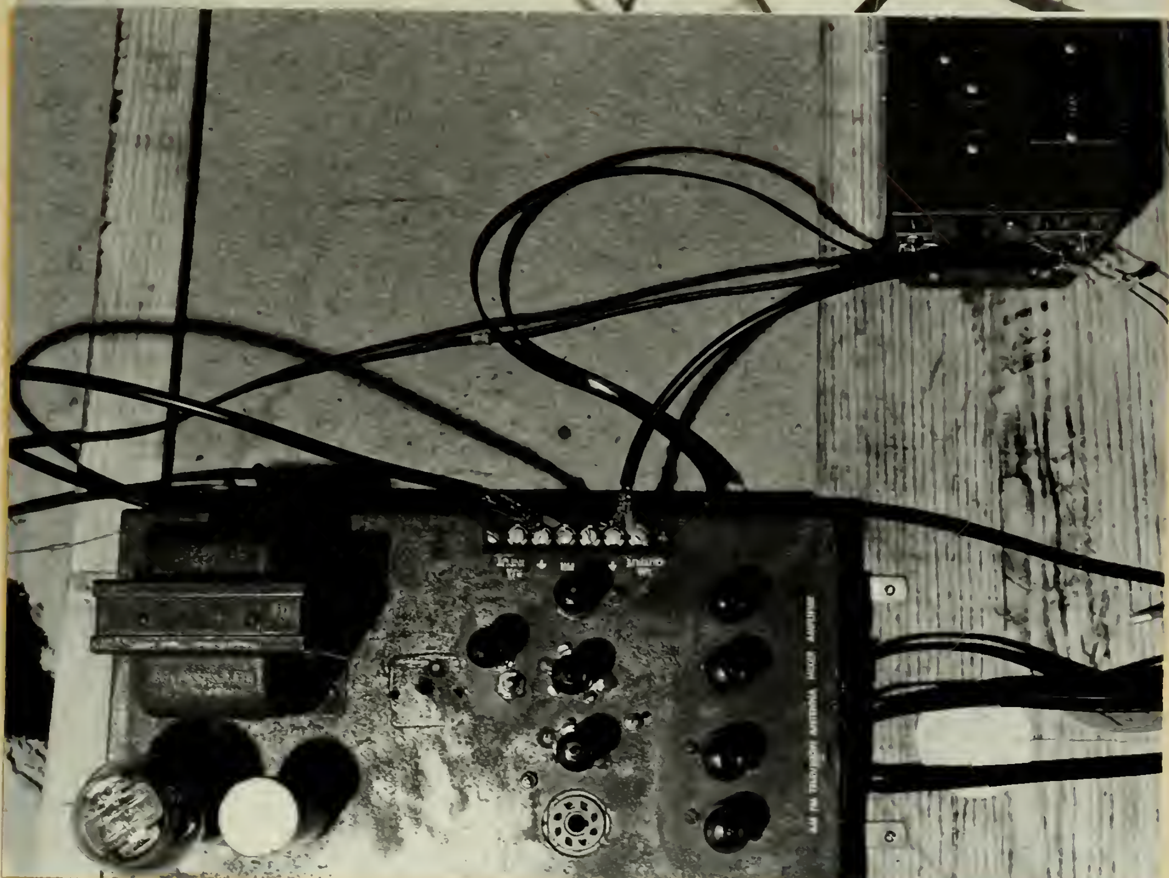


Figure 25(a)
Simplified Circuit



is approximately 150 ohm to cover the VHF Television range, triodes are used in pairs to feed from and distribute to 300 ohm balanced lines.

Figure (26) shows photographs of an 11 outlet distributed line distribution unit built at the U. S. Naval Postgraduate School for use in the B. O. Q. Figure (27) shows a wide band booster under a 4 outlet distribution unit built at the U. S. Naval Postgraduate School for use in the B. O. Q.

Figure (25) shows the IT-102A auto-booster which gives 19 db of gain with a 9 db noise figure.

While the similarity of this unit to a group of pi couplings is hard to see, the circuit actually is a development of the one in figure (25(b)) which ignores voltage feeds and isolating RF chokes. Series Peaking circuits are added to flatten response, and the R. F. chokes are made resonant at some mid-frequency to boost gain. Using grounded grid tubes throughout reduces the noise to an acceptable level. The Blonder-Tongue Company makes a similar unit in which only the first tube is grounded grid. The noise figure is almost doubled.

CHAPTER V

INTERSTAGE COUPLING

The use of pi sections as impedance matching sections is not a recent innovation. It is, in fact, almost as old as the art itself. Several of the earlier circuits are shown in figure (28).

In post World War II television circuits T and pi networks are used to couple I. F. stages, video stages, and more recently R. F. stages in the form of cascade or driven grounded grid R. F. amplifiers and boosters.

Perhaps the best example of how the pi coupling has developed is the driven grounded grid. Figure (29(a)) shows the circuit as originally used by television manufacturers.

Figure (29(a)) shows the basic driven grounded grid circuit using a special dual triode such as the 6BQ7, 6BK7 or 6BZ7. An improvement is shown in 29(b). The inductance L_1 serves two purposes, first, to resonate with the plate and cathode capacitances and thus match impedances; second, to cause a 180° phase shift between plate and cathode, thus giving almost perfect neutralization because input and output are 180° out of phase and any feedback is negative, this results in 1 or 2 db improvement in noise figure. Figure (29(c)) shows a further refinement for higher gain with good neutralization. The tapped inductance L_1 matches a low impedance cathode (150 ohms) to a plate at about 1500 ohms with a modified pi match. With the plate driving the cathode directly, the first stage has a gain of approxi-

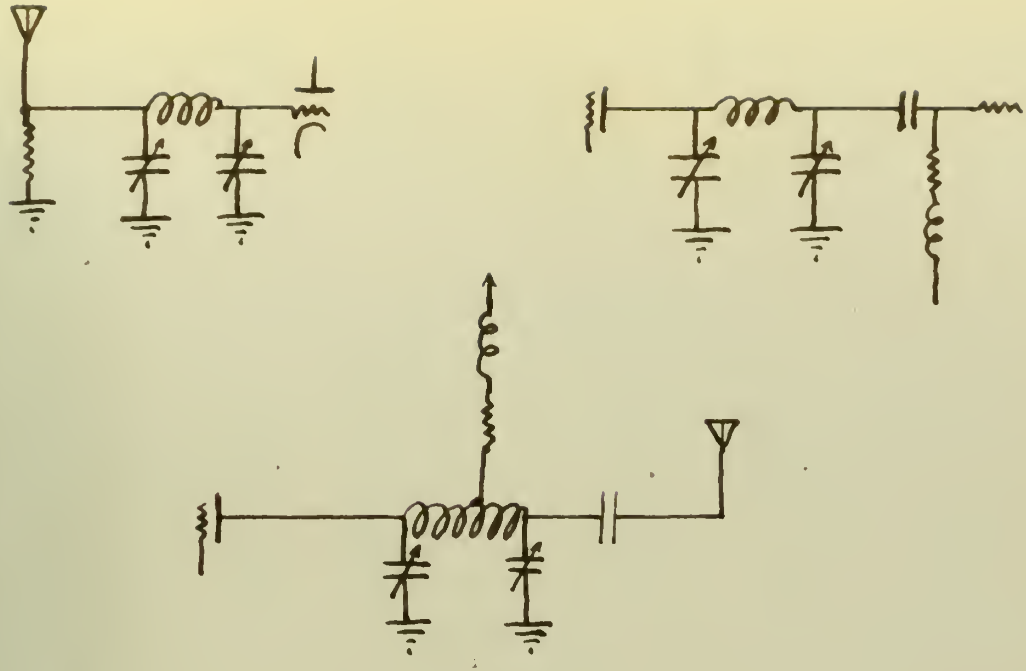
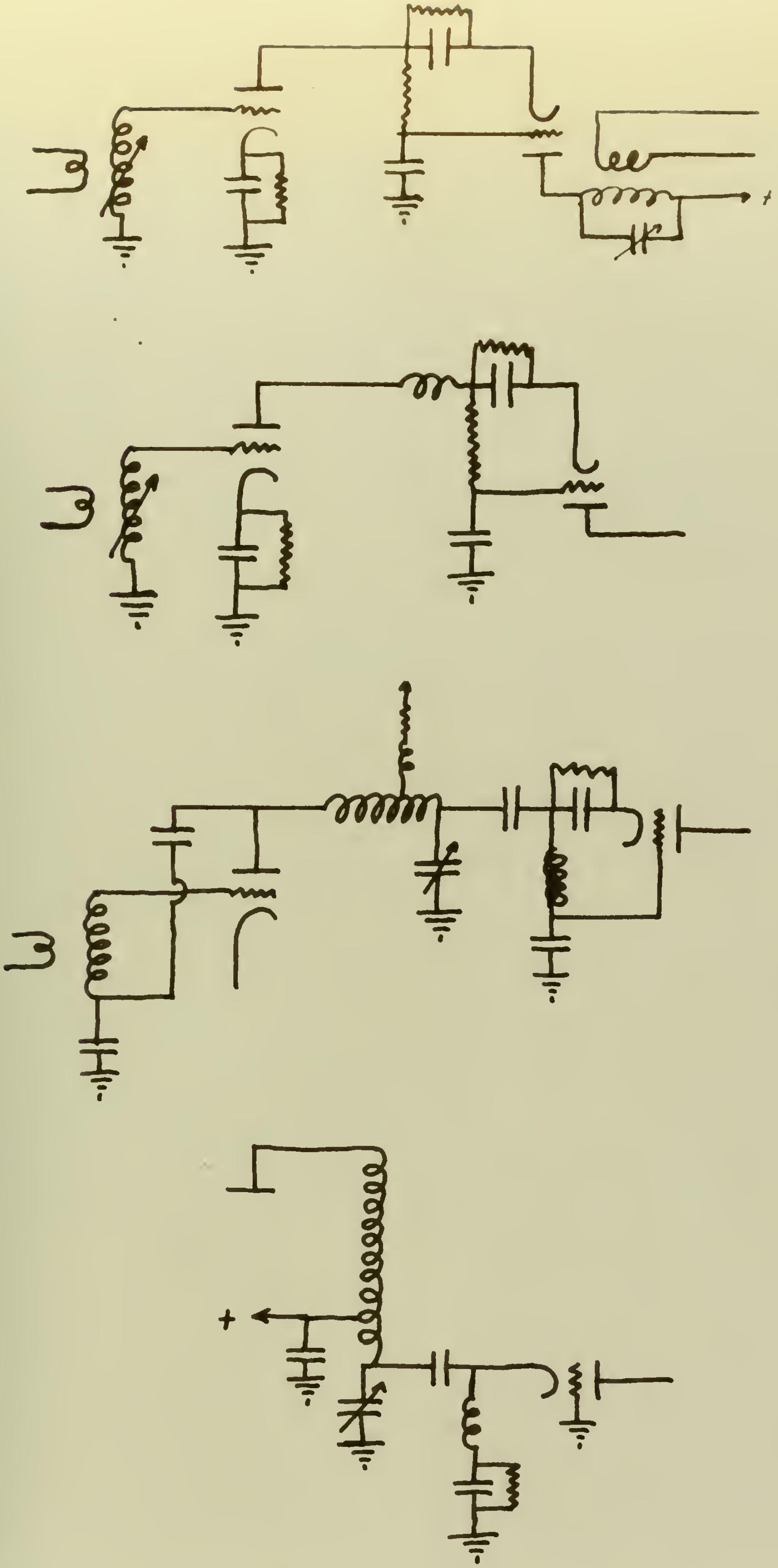


Figure 28 – Matching Sections (π)



(a)

(b)

(c)

(d)

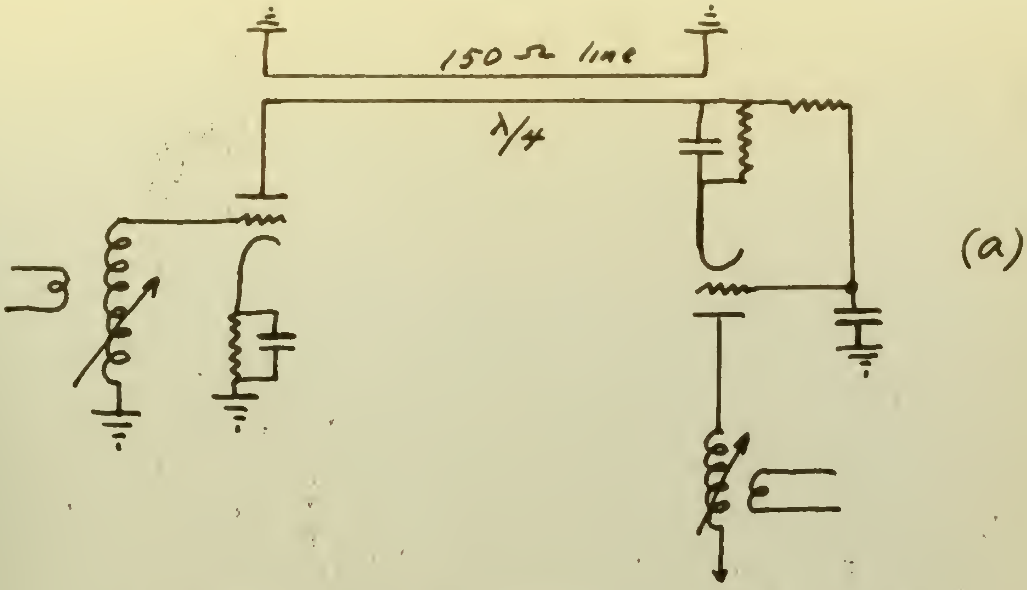
Figure 29 - Cascode Circuits

mately 1. Using the auto transformer L_1 the gain is $\sqrt{\frac{1500}{150}} =$ 3.2 or approximately 10 db. The output phase is still negative so that good neutralization is achieved. L_2 is used to resonate with the cathode to obtain a pure resistive load to the output of L_1.

An interesting variation of this circuit used by Philco (52) uses an actual transmission line section of 150 ohms twinlead. This circuit is shown in figure (30). The plate sees the end of a transmission line loaded with its Z_0, hence the load on the plate is 150 ohms. For any feedback signal however, the plate is essentially an open circuit, and the cathode sees a short. In this way better neutralization is achieved.

The pi section may be used in IF stages to obtain a relatively flat response through the I.F. region, but rejecting the fundamental R. F. and oscillator frequencies. The circuit shown in figure (30(b)) is used by Admiral (52) to couple the mixer pentode to the first I. F. by loading the plate with a small additional condenser is obtained, and the resulting pi has a Q of 2 or 3 with a sharp cutoff at about 30 mc.

Artificial line sections are not frequently used as I.F. interstage couplings, most manufacturers preferring stagger tuned circuits with traps or M derived filter sections.



$$Z_o = R_L = \frac{1}{G_m} = 150\,\Omega$$

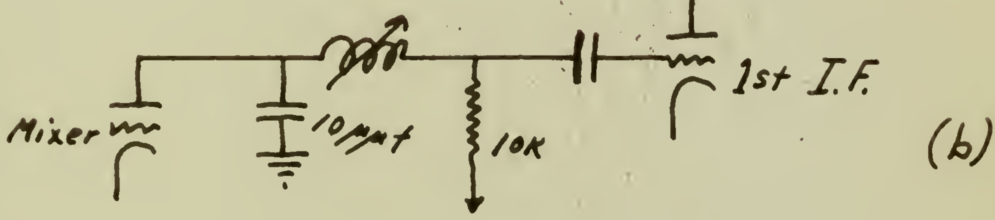


Figure 30
Tuner And Mixer Coupler Circuits Used By
Admiral and Philco

CHAPTER VI

THE VIDEO AMPLIFIER

Compensation of a video amplifier to get flat response may be treated as an extension of the basic line theory. Figure (31(a)) shows the video amplifier compensation as normally drawn. Figure (31(b)) shows the compensation redrawn as a ladder network of two L sections.

The capacitors C_1 and C_2 are tube output and input capacitors. It is interesting to note that the feed point may be shifted to the center of the ladder if plate capacity is smaller than grid capacity as in figure (31(c)).

The simple case of series peaking is merely one of adjusting a pi network for maximum flat response. The plate supply resistor is placed on the side where C is smallest, in this way the circuit Q is controlled at high frequencies with maximum gain at lower frequencies. See figure (31(d)).

The method of design is to find the 3 db down point of the uncompensated amplifier, i.e., the frequency at which total capacitive reactance equals the plate load resistance, then pick an L to resonate at 1.4 times this frequency at a Q of 1 with R_1 as the shunting Q limiting factor and with C_1 and C_2 in series. Obviously this requires some juggling of values to obtain, but when once adjusted, a very good response curve is obtained.

It should be stated here that while the series peaking and combination peaking circuits give higher gain, they are in disrepute in high quality systems because their sharp cutoff tends to cause ringing by

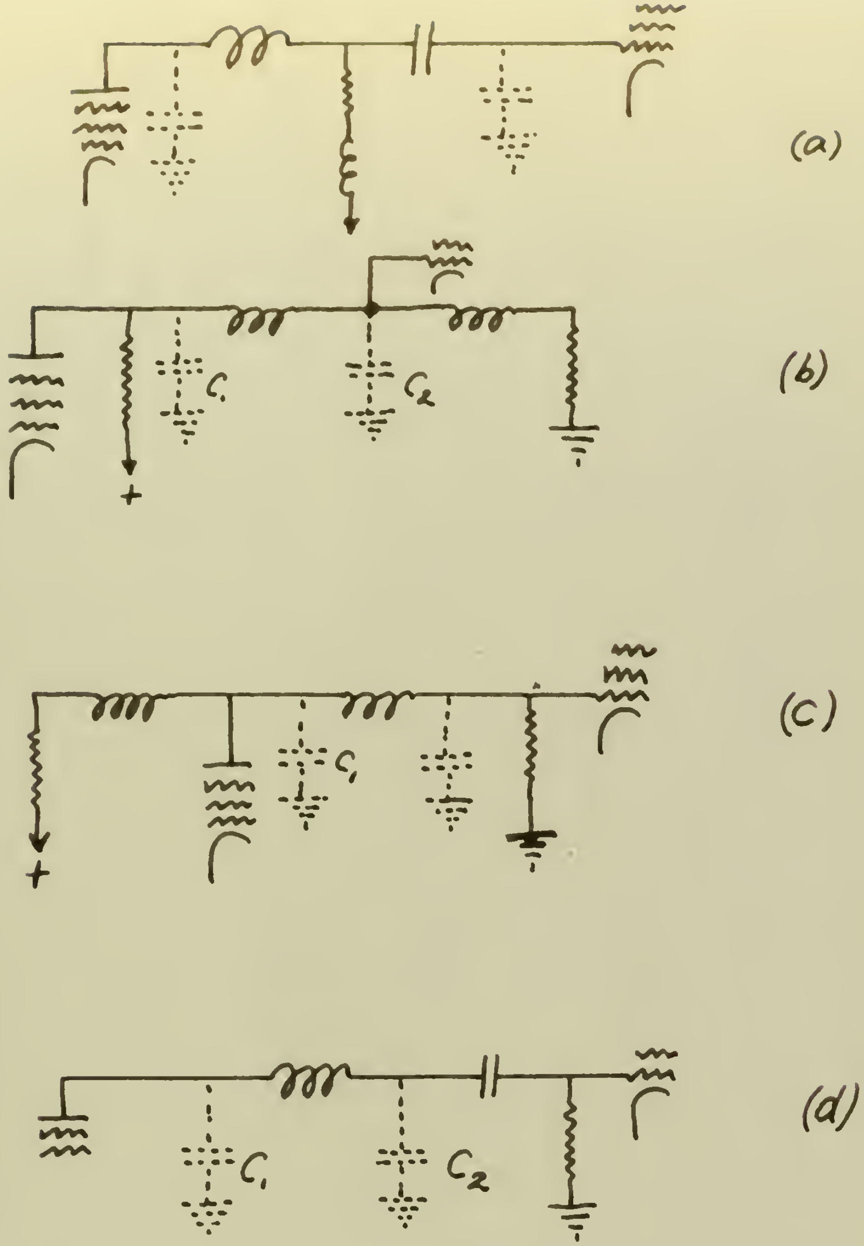


Figure 31 - Video Coupling Circuits

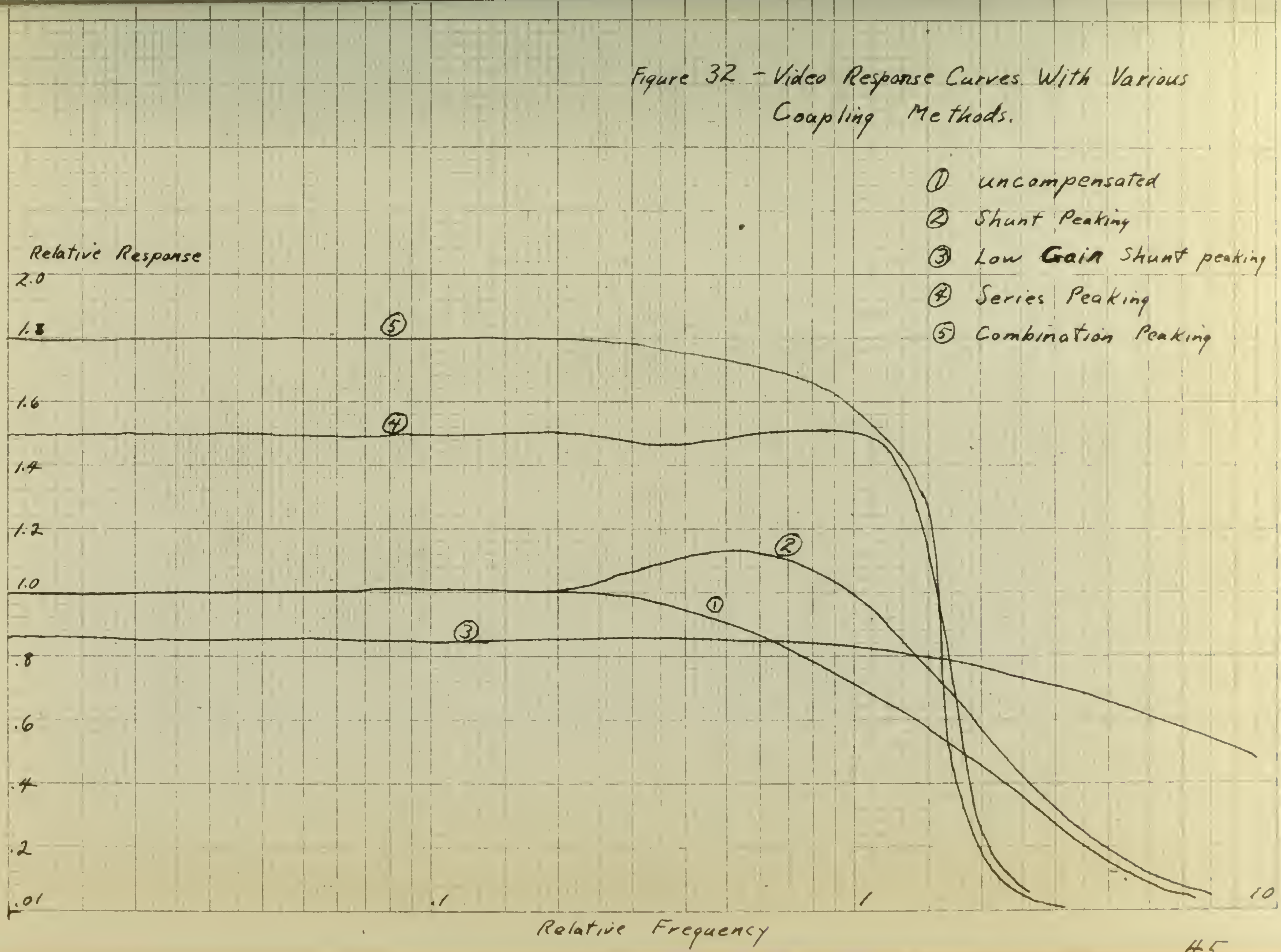


Figure 32 - Video Response Curves With Various Coupling Methods.

transients and consequent distortion of the picture.[37][39] Combination and series peaking are little used in transmitters, and find their chief application in receivers where the cutting off has already been done by transmitting diplexers and receiver I. F. stages. Shunt peaking is to be preferred in television circuits in spite of the lower gains achieved.[37][39]

Methods for calculating the optimum values for these video peaking systems have been worked out,[57] and the rules for determining optimum values are found in nearly all texts,[53][54][55] so that with the possible exception of the pi match or series peaking, little or no advantage is obtained by using the methods set forth in this thesis for video compensation. Figure (32) shows the relative responses of the various methods of video compensation.

CHAPTER VII

EXPERIMENTAL CIRCUITS

The use of pi sections presents several interesting possibilities in R. F. circuits. One of these is the possibility of achieving an extremely low noise figure for I. F. stages coupled to a balanced mixer, or for television R. F. stages coupled direct to balanced inputs. Figure (33) shows the basic circuit of the low noise input devised by the author.

In determining noise figure, the actual circuit resistance is compared with a theoretically perfect tube so that the actual input resistance divided by the ideal input resistance yields have equation (21). Figure (34) shows the equivalent circuits of the actual and ideal cases. The tube is considered ideal.(33)(34)(43)

If R_1 is made large, the effect of noise resistance is decreased, i.e., if we use a high Q circuit at the input, the noise figure will decrease with increasing Q. A point is soon reached, however, in television circuits where the bandwidth becomes too narrow, or tube input conductance limits Q by its shunt resistance effect.

The circuit of figure (33) uses 2 pi coupling units at the input which are higher than normal Q for the bandpass required and stagger tuned. Looking in at the low impedance end, the Q is very low and the feed system sees a balanced input, but at the grid side the Q is high and the input is thus frequency selective. As an example of what might be done with an input of this type, assume a signal is desired at 69 mc (channel 4) with a 6 mc bandpass. To do this with a conventional input,

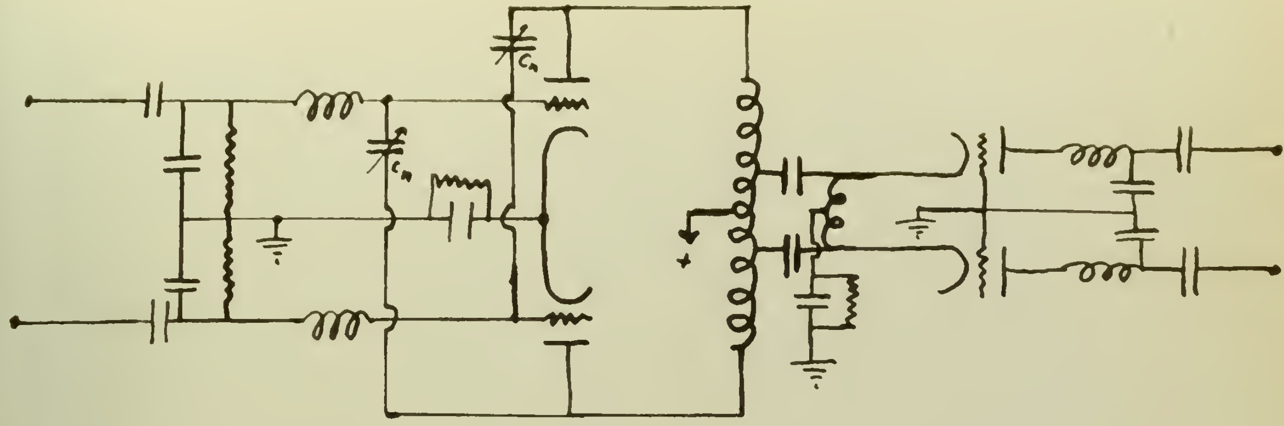
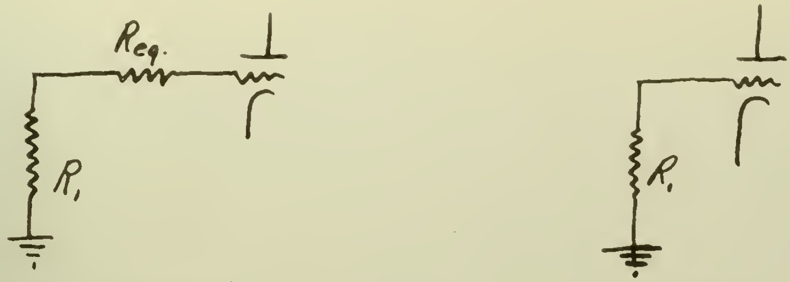


Figure 33 - A Low Noise R.F. Amplifier



actual Circuit　　　　Ideal Circuit

$$E_n = K\sqrt{R_{eq}}$$
$$= \sqrt{4kT\Delta f}\,\sqrt{R_{eq.}} \qquad (20)$$

$$n.f. = \frac{R_1 + R_{eq.}}{R_1} \qquad (21)$$

Figure 34 Actual and Equivalent Circuits For Determining Noise Figure.

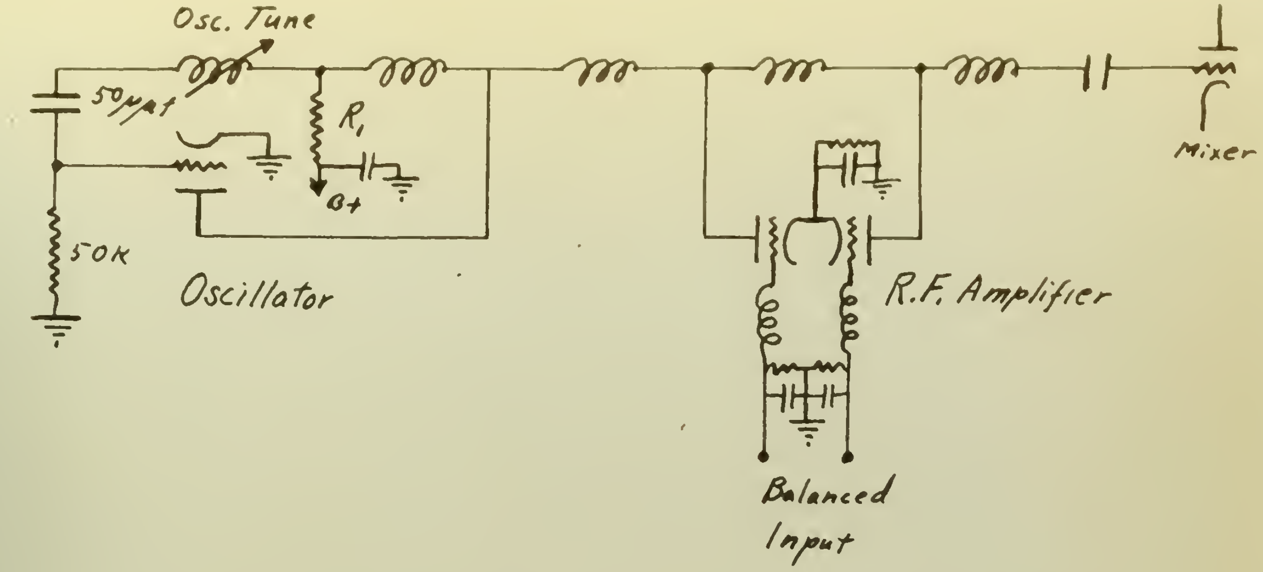


Figure 35 - A Distributed Line Converter.

the Q cannot exceed 10 or 11. Using stagger tuned pi units, one tuned to 67.5 and the other to 70.5, the Q can be made equal to 20 and a 6 mc bandpass, with a flatter top characteristic can be maintained. In so doing, the resistance R_1 is raised and the effect of R_{eq} is decreased. An improvement in noise figure up to 3 db is possible.

A unit to test this circuit was built by the author and tested in January of 1953 using 2-6BK7 tubes. Making measurements with a Kay Mega-node a noise figure of 3 db was obtained.

Another interesting possibility is the use of a long line section for oscillator tank, mixer, and R. F. stage. Figure (35) shows the circuit devised by the author.

In figure (35) the inductances and tube elements form a line which has a relatively high Q at the oscillator, and R. F. The mixer grid being a mismatch focal point equivalent to an open ended line. Standing waves are created by R_1 which is smaller than Z_o.

Such a circuit has little to recommend it at lower frequencies, but at the very high or U.H.F. regions the use of simple loops or hairpin inductances with slide tuners can result in a very desirable unit, chiefly because no inductive coupling is required and many components such as R. F. chokes and by-pass condensers are unnecessary.

The oscillator will oscillate at the frequency at which its interelectrode inductance gives a 180° phase shift. The Q of the oscillator tank is determined by the tuning of the plate sections and R_2. The mixer has maximum gain at the frequency at which its interplate inductance has a 180° phase shift. Both sections pass their outputs down the artificial

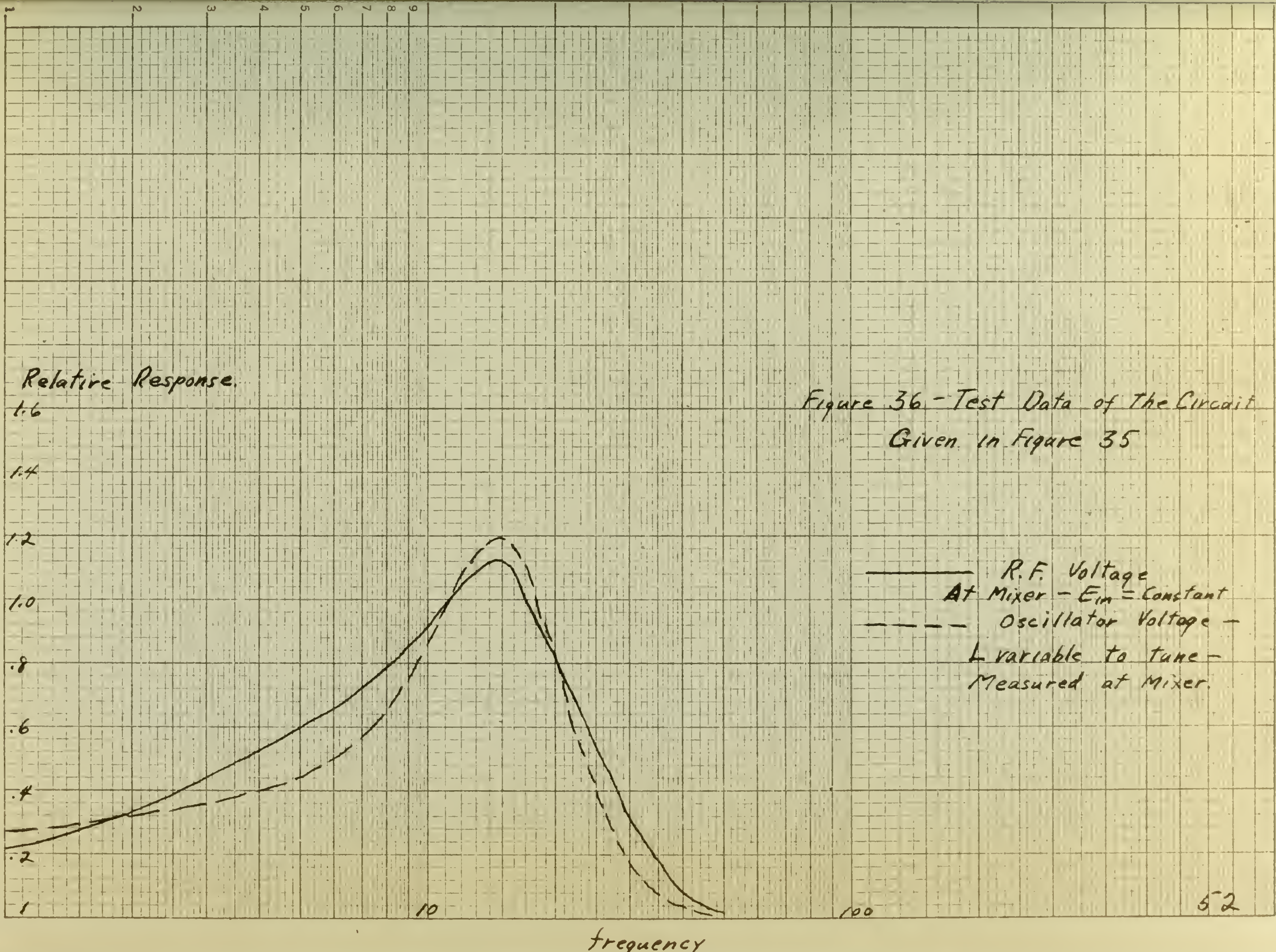


Figure 36 – Test Data of The Circuit Given in Figure 35

line so that they tend to reflect from end to end causing a standing wave at the mixer.

The circuit of figure (35) has been built and tested. Results of test data and curves are given here in figure (36). In order to get the most reliable data in the easiest way, a technique which might be called frequency scaling is used. Large inductances and capacities are employed so that the system which would normally be operating in megacycles can be tested in the kilocycle range and such factors as amplitude and phase can be observed on an oscilloscope.

Several other circuits employing line sections are given in the Appendix without explanation or comment.

APPENDIX A

Examples of Applications of Figure 3

I. Plate to Grid Coupling

A given plate circuit has 10 $\mu\mu$ f of capacity which is 1000 ohms X_c at the operating frequency. The following grid has also 10 $\mu\mu$ f of capacity. The bandpass requirements are such that a Q = 10 will serve. Since a condenser input and output is present, pi coupling is indicated. The Q = 10 is large enough that X_L can be made equal to X_c, and R_2 will equal 10 X_c 10,000 ohms. R_1 will equal 100 ohms. Referring to the pi equivalent in figure (4), R_1 is an internal impedance at the center and is not an actual resistor. R_2 must be an actual resistor at one end, to supply B + to the plate. Since there are 2L sections to make the pi, the L chosen will be twice the value indicated above, or X_L 2000 ohms. The final design is shown in figure (5). This is an actual mixer to I. F. coupling used by Admiral.

II. Line to Grid Coupling, or plate to line

It is desired to connect a 6AK5 grid to a 75 ohms line. The input capacity is 11 $\mu\mu$ f. The frequency range must be as broad as possible, preferably reaching 200 mc. Referring to reactance charts figure (6) we find 11 $\mu\mu$ f has a reactance of 150 ohms at 100 mc, or 75 ohms at 200 mc. We can make a one to one match with Q equal zero at 200 mc. Referring to figure (3), we note the ratio of X_c to X_L for Q equal 0 is infinite. An arbitrary decision must therefore be made. Since the loading resistance is the line, the phase can be kept near zero shift if we make

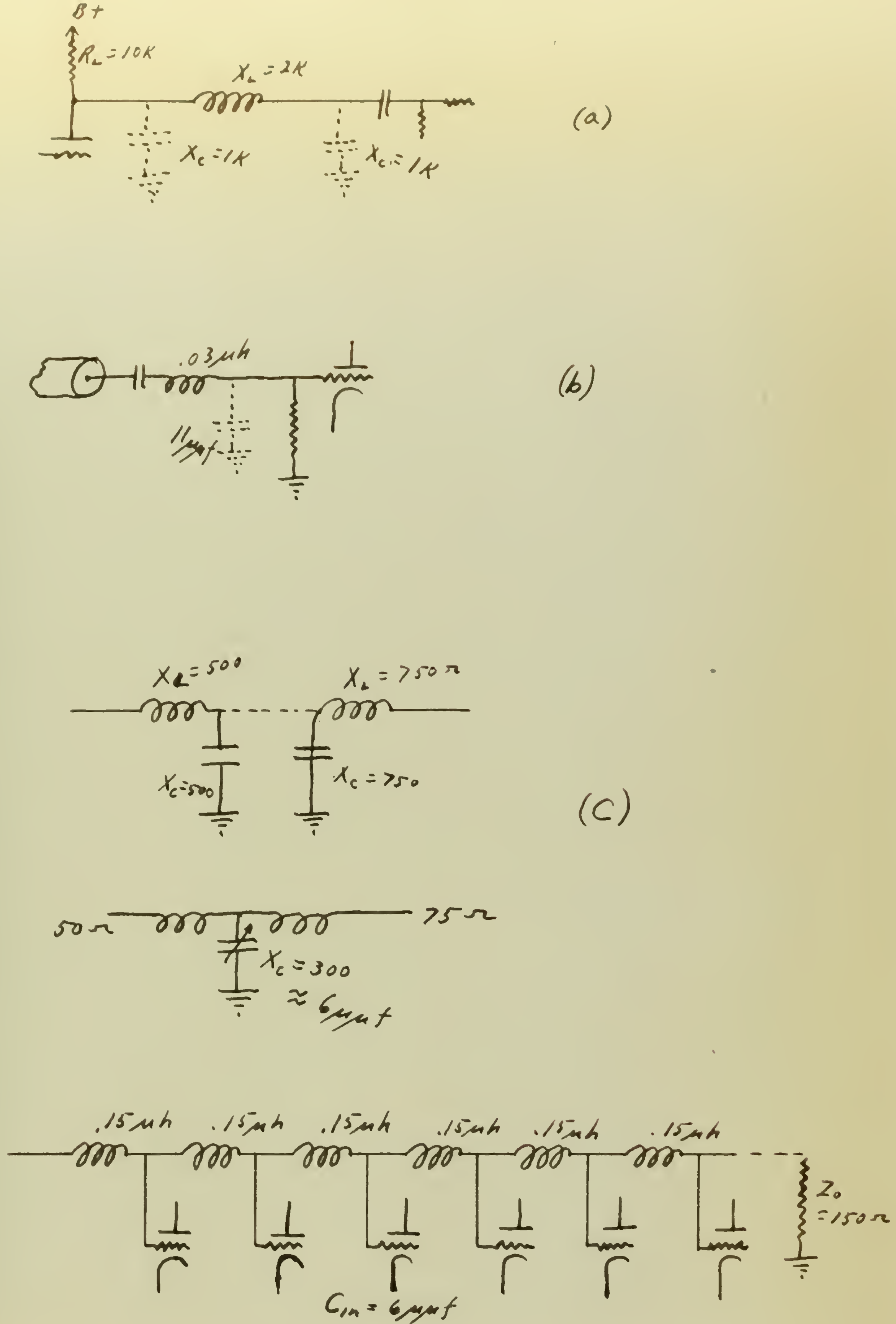


Figure 5 - Examples of Impedance Matches.

$X_C = X_L$ at 1.4 times 200 mc or 280 mc. At this frequency X_C = 53 ohms, so X_L is picked as 53 ohms at 280 mc, or 38 ohms at 200 mc, or 19 ohms at 100 mc, which corresponds to .03 microhenry. The same approach can be used for plate to line coupling. See figure 5(b).

III. Line to Line Coupling

It is desired to match a 50 ohms line to a 75 ohms line with a frequency selective match. An overall Q of 10 at 100 mc. is desired. A T match is indicated. As was shown in Chapter II, the Q will stay 10 if two Q = 10 sections are linked. If Q = 10, R_1 = 50 ohms, X_C = 500 ohms. Since Q is large X_L will also equal 500 ohms. R_2, which is an internal impedance and not an actual resistor or load is 5000 ohms. These values can easily be picked off the chart in figure 3. Since the T contains two L sections joined at R_2, the actual value of C would be the two values paralleled. Such a circuit would have to be tuned for best results. See figure 5(c).

IV. Line to Multiplate Stage Coupling

Ten grids must be paralleled to be driven from a 150 ohms line. the 12AV7 which has a total input capacity without Miller effect at 6 is to be used. Ten grids at 6 would give a total input capacity of 60 if paralleled. If an artificial line with Z_0 = 150 ohms is created, the input will be 150 ohms of pure resistance. Such a line must be terminated in Z_0 if it is to be flat. There are several approaches to getting values for L, but since L/C must equal $(Z_0)^2$ we can merely solve

for this relationship, or refer to the chart-figure (6) and pick the values off directly. The intersection of 6 $\mu\mu f$ with 150 ohms is also the intersection of .15 μh .

The use of line matching sections is not limited to the purely resistive cases. For narrow frequency bands or single frequencies, it is permissable to use the reactive element to be matched as part of the matching section.[(15)] Thus, if a capacitive reactance is present use it in the L section instead of an external C, or reduce the external C accordingly. If the reactance is inductive, subtract it from the value of X_L to be used in the match. Using this method it is possible to match antennas with widely varying complex impedances to a line or tube.

Division of Complex Numbers

Appendix B

Let Z_1 = Complex no. 1

Z_2 = Complex no. 2

by similar Triangles

$OZ_1'Z_2'$ similar to OZ_2Z_1

but $OZ_1' = 1$

$$\frac{OZ_1'}{OZ_1} = \frac{OZ_2'}{OZ_2}$$

$$\frac{OZ_1'}{OZ_2'} = \frac{OZ_1}{OZ_2}$$

$$\frac{OZ_2'}{1} = \frac{OZ^2}{OZ_1}$$

Normallized

$$|Z_2'| = \frac{Z_2}{Z_1}$$

$$\overline{Z_2'} = Z_2' e^{j(\alpha_1 + \alpha_2)}$$

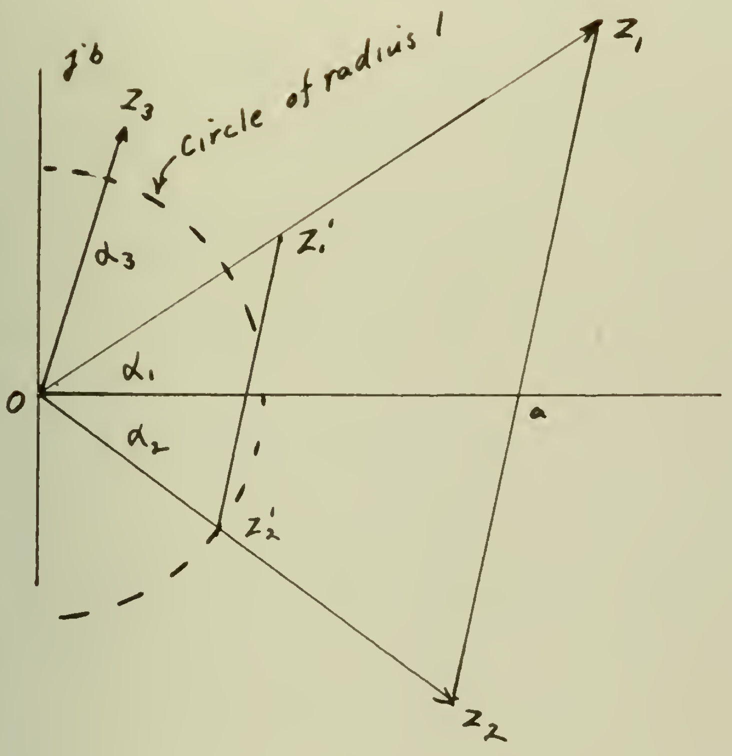


Figure 37

$$G(s) = \frac{R_2}{\left(\frac{L_1}{C} R_2\right) s^2 + \left(\frac{R_2^2}{C} + \frac{L_1}{C^2}\right) s + \frac{R_2^2}{C^2}} = \frac{E_2(s)}{E_1(s)}$$

Let $E_2(s)$ be 1 (see Fig. 13)

Then

$$E_1(s) R_2 = \left(\frac{L_1}{C} R_2\right) s^2 + \left(\frac{R_2^2}{C} + \frac{L_1}{C^2}\right) s + \frac{R_2^2}{C^2}$$

$$E_1(s) = \left(\frac{L_1}{C}\right) s^2 + \left(\frac{R_2}{C} + \frac{L_1}{C^2 R_2}\right) s + \frac{R_2}{C^2}$$

$$E_1(s) = A s^2 + B s + C$$

$$E_1(s) - Bs - C = A s^2$$

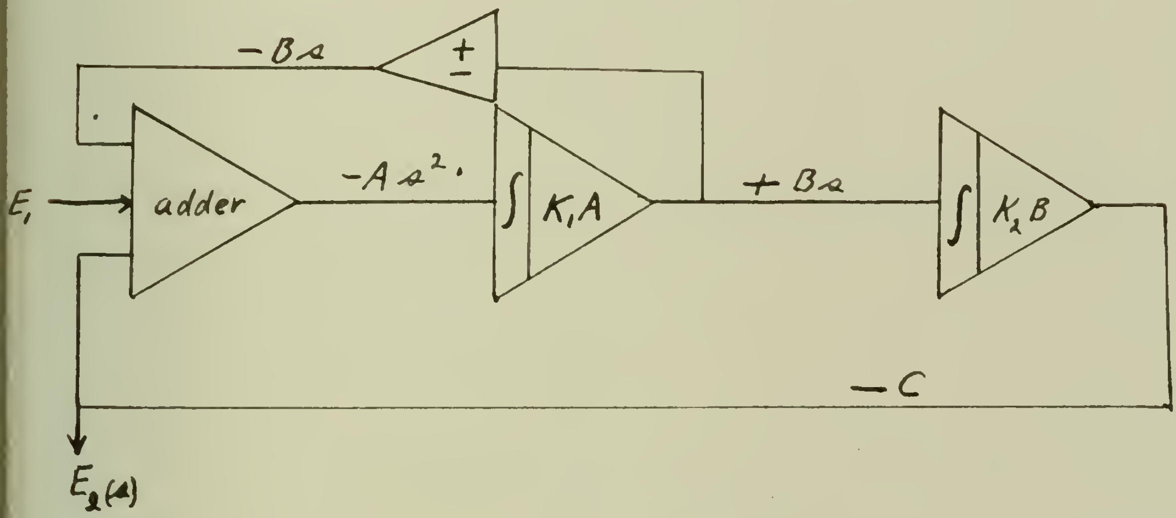


Figure 38 - Analog Computer Data For the L Match

Let $E_3(s) = 1$ (see Fig. 15)

$$s E_1(s) - Bs^2 - Cs - 1 = As^3$$

$$E_1(s) - Bs - C - \frac{1}{s} = As^2$$

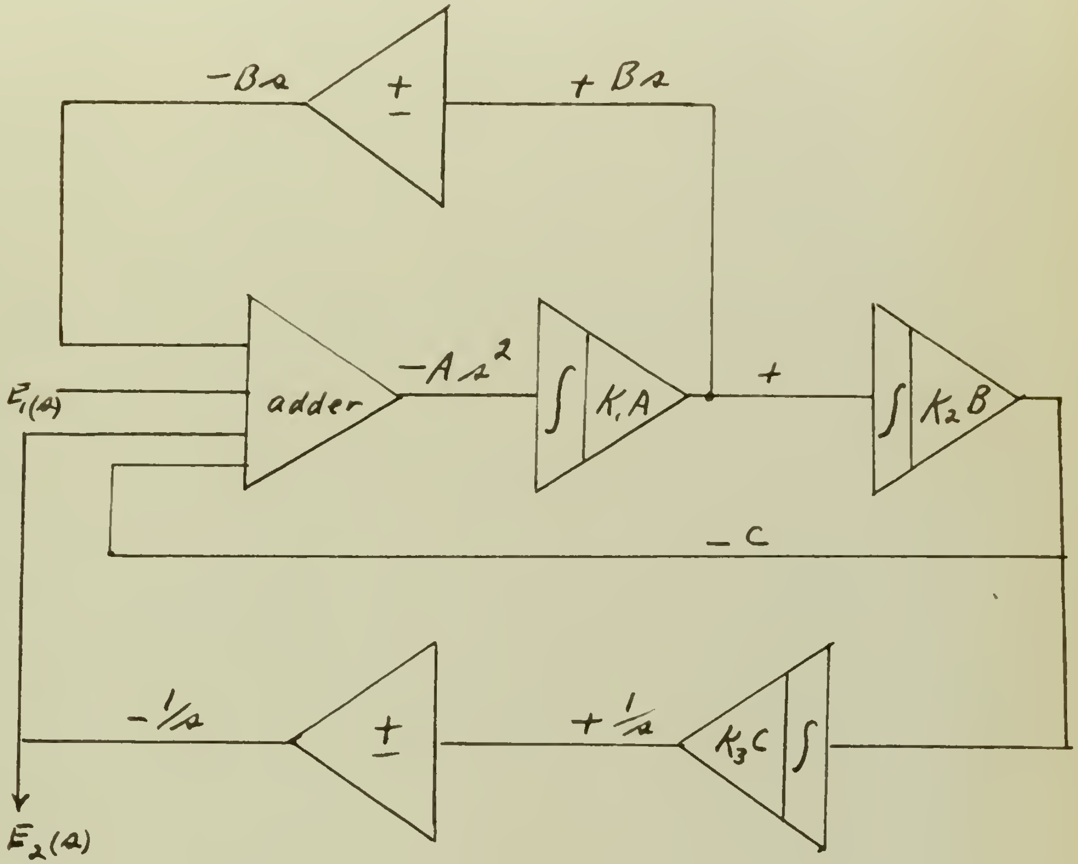


Figure 39 Analog Computer Data For T Match

$$E_1(s) - Bs^2 - Cs - 1 = As^3$$

(See Fig 14)

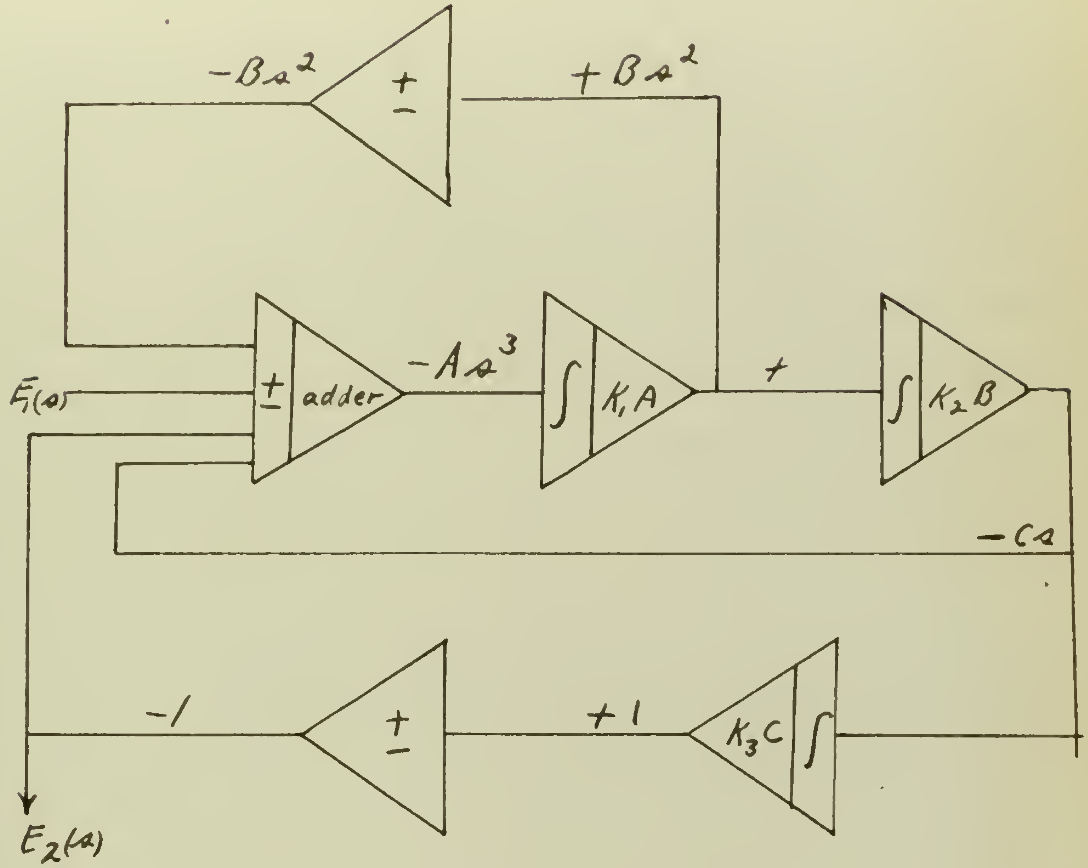


Figure 40 Analog Computer Data For π Match.

APPENDIX D

MASTER CAMERA - RECEIVER SYSTEM FOR INDUSTRIAL TELEVISION

The Radio Corporation of America required a master system for use in a customer's plant which would permit an Industrial Television camera to be plugged in at any point, and any number of receivers to be plugged in at other points without interference. It was stipulated that no more than twelve (12) cameras would be operating simultaneously, and each of these could be attached to a small R.F. unit with approximately 2 Volts output feeding a 75 ohm coaxial line. The following solution was proposed by the author, who was then attached to The Radio Corporation of America.

Resistor type drops are used, there being no reason to conserve signal strength. A total athenuation of 50 db is obtained, plus line loss, which still leaves a signal of several millivolts at each receiver.

Loss of signal between points

$$A = 2\gamma l$$

Voltage Reflected Ratio $|K| = \frac{\rho - 1}{\rho + 1} = \frac{Z_n - Z_0}{Z_n + Z_0}$

$$\rho = \frac{Z_0}{Z_n}$$

A slight discontinuity is to be created by Tapping a coax feed line with a high resistance takeoff to a receiver — 40db of ghoust rejection is required.

If $2\gamma l = 0$

$$-40\ db = \frac{1}{100} \qquad \rho \approx .98$$

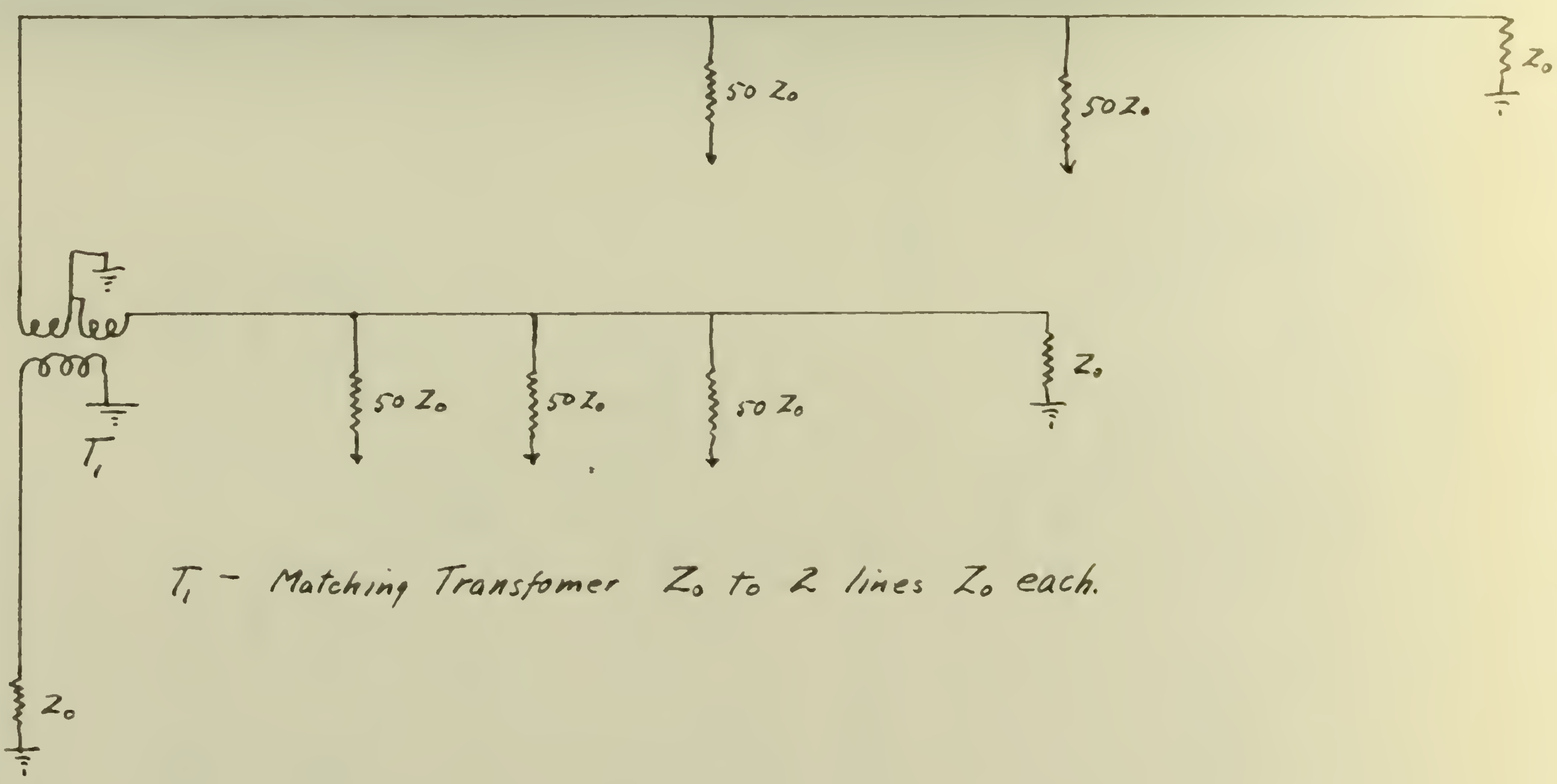
$$|K| = \frac{1}{100} \approx \frac{.98 - 1}{.98 + 1} = \frac{.02}{1.98} \approx -40\ db$$

for $\rho = .98$ Z_n must be $50 \times Z_0$

for 75Ω coax $Z_n = 3{,}750\Omega$ min.
50Ω coax $Z_n = 2{,}500\ \Omega$ min.

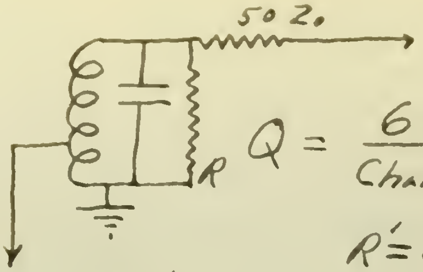
If $2\gamma l \neq 0$, Z_n may be decreased, or as is done in practice — more outlets can be added.

Figure 41 — Development of maximum Standing Wave Permissible on a master Antenna System.



T_1 - Matching Transfomer Z_0 to 2 lines Z_0 each.

Figure 42 - Master Camera-Receiver Distribution System for Industrial T.V. and/or master Antenna System with any number of outlets and camera inputs.

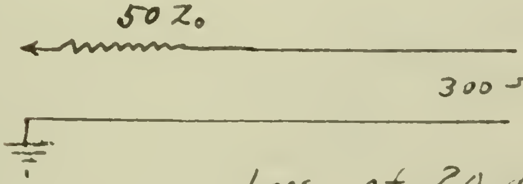


To coax distribution line.

$$Q = \frac{6\,mc}{\text{Channel Freq.}}$$

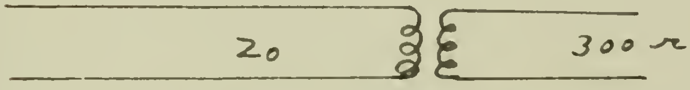
$R' = Q X_c$ = parallel equivalent of $R + (Z_0 + 50 Z_0)$

Coax feed from ITV Camera or antenna system boosters. Provides Matched load to signal Source.



300 Ω to receiver

Loss of 20 db approximately is obtained using this type of coupling. Signal on line must be great enough to make up this loss.



Transformer Coupling for Use with Receiver Used as line Terminus. This is not Recommended Because Receivers Seldom Match at Less than 1.25 VSWR

Figure 43 - Miscellaneous Circuit For Use with Master Antenna Systems.

RCA - AM-FM-T.V. Mixer

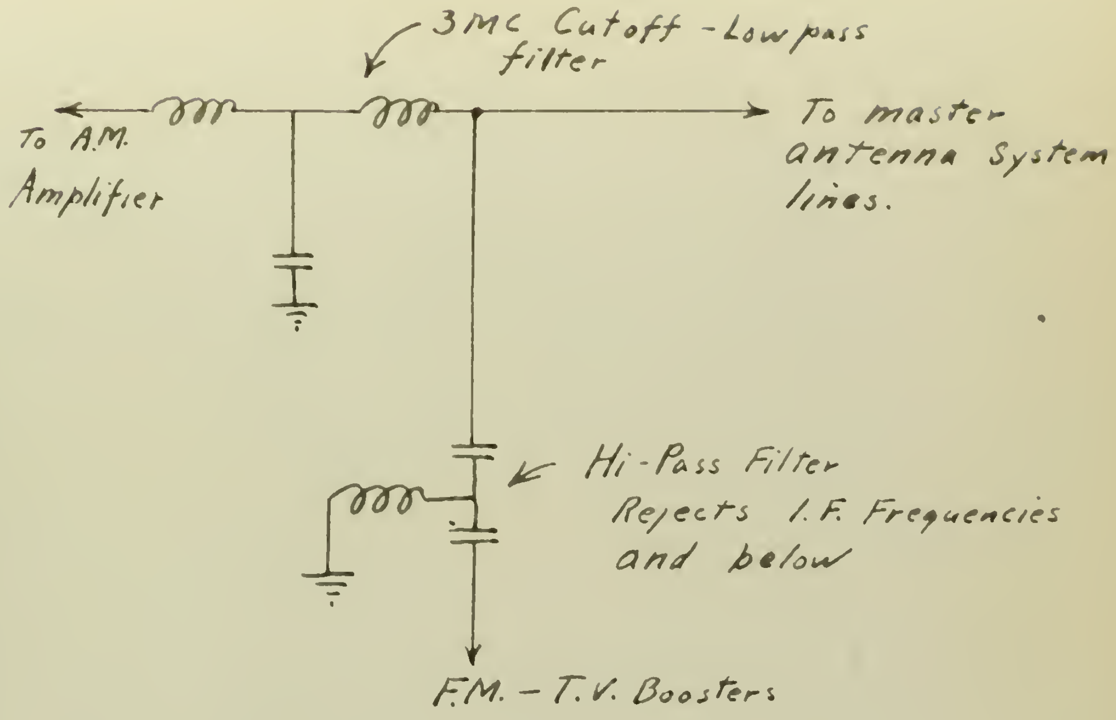
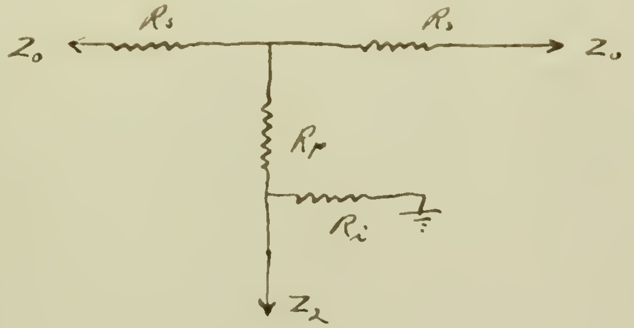


Figure 44 - Miscellaneous Circuit For Use With Master Antenna Systems.

T Pad to be inserted in Coax Line for perfect match.



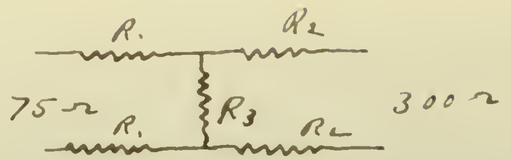
$$Z_2 = \frac{R_i\left(R_p + Z_1 + R_s/2\right)}{R_i + R_p + Z_2/2 + \frac{R_s}{2}}$$

or to a close approximation

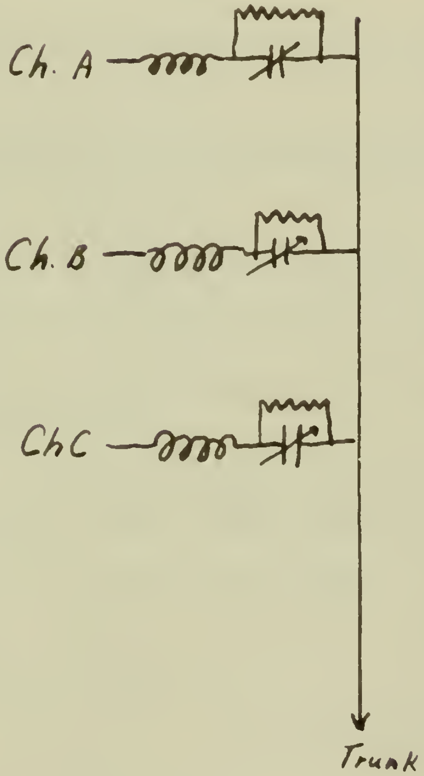
$$R_i = \frac{R_p Z_2}{R_p - Z_2}$$

Figure 45

Resistor Pads



Loss db	R_1	R_2	R_3
12 db	0 Ω	130	82
15 db	13	130	56
20 db	24	130	30
25 db	30	130	18



Lynmar Signal To Trunk Mixer Feed System.

Figure 46

BIBLIOGRAPHY

1. The Distributed Pair - Pederson - Professional Group on Circuit Analysis - I.R.E., Dec. 52 (PGCT1).

2. Impedance Admittance Conversion Chart - Paine - Electronics - Jan. 46, pg. 162.

3. Kahal - Transactions A.I.E.E. - Synthesis of Four Terminal Networks - From Doctoral Dissertation.

4. Darlington, S. - Journal of Mathematics and Physics, Vol. 18, No. 4, Sept. 39.

5. Impedance Combining Chart - Muffley - Electronics - March 44, pg. 134.

6. Impedance Conversion Chart - Series to Parallel - Ware Electronics - Oct. 44, Pg. 144.

7. Impedance, Magnitude and Phase Angle Charts - Blow - Electronics - Jan. 43.

8. Impedance, Magnitude and Phase Shift Curves for Some Common Two-Terminal, Three Element Linear Networks - Edutis - Electronics - Nov. 42, pg. 76.

9. Impedance Measurements at V.H.F. - Hills - Dec. 45, pg. 156.

10. Impedance Nomograph - Wheeler - Electronics - Sept. 46.

11. Impedance Plotter - Raymond and Drumheller - Electronics - March 49, pg. 128.

12. Paine - Geometric Solutions for Resonant Impedance Matching Networks - Radio - Oct. 43, pg. 36.

13. Bruene - Pi Network Calculator - Electronics, May 45.

14. Storch - Design Procedure for Pi Network Antenna Couplers - Proc. I.R.E. - Dec. 49.

15. How to Design R. F. Coupling Circuits - Bruene - Electronics - May 52, pg. 134.

16. Impedances, Equivalent Series and Parallel - Peek - Electronics - June 45, pg. 254D.

17. Transmission Line Calculator - Smith - Electronics - Jan. 44, pg. 130 and Jan. 39, pg. 29.

18. Network Design Charts - Foley Electronics - Sept. 51, pg. 132.

19. L Type Networks - Minimum Loss - Electronics - Feb. 51, pg. 118.

20. Impedance Matching Theoretical Limitations - Electronics - Feb. 51, pg. 234.

21. Bandwidth of Quarter Wave Sections - Fubini - Electronics - Oct. 52, pg. 138.

22. R. F. Coupling Nomograph - Electronics - Oct. 52, pg.166.

23. Impedance Circle Diagrams - Radio and Television News - Engineering Edition - Feb. 51, pg.15.

24. Matching Stub Calculations - Radio and Television News - Engineering Section - Aug. 51, pg. 31.

25. Length of Transmission Line Nomograph - Radio and Television News - Engineering Section - Feb. 50, pg. 32.

26. Adaptation - Sur Une Large Bande de Frequences par Plusieurs Organes Intermediares - Aberdam - Onde Electrique - Dec. 48, pg. 474.

27. Antennas Input Systems - Foster - Tele-Tech. Feb. 49, pg. 28.

28. Resonance Characteristics by Conformal Mapping - Honnell & Horn - Proc. I.R.E. - Oct. 52, pg. 1211.

29. Transmission of F. M. thru a Network - Frantz - Proc. I.R.E. March 46.

30. Input Impedance of Several Receiving Tubes - R.C.A. Licensee Bulletin LB661 - March 46.

31. An M Type Band Pass T.V. Amp - R.C.A. Licensee Bulletin - LB687 - November 46.

32. Input Admittance of Receiving Tubes - R. C. A. Application Notes AN-118.

33. Some Notes on Noice Theory and its Application to Circuit Design - R.C.A. Review - Sept. 48.

34. Noise Factor and its Measurement - R. C. A. Licensee Bulletin LB775.

35. Resonant Frequencies and Characteristics of a Resonant Coupled Circuit - R.C.A. Review - Sept. 50.

36. Standing Waves and Impedance Circle Diagram - Westcott - Wireless Engineer - July 49.

37. Transient Response of Filters - Corrington - R.C.A. Review - Sept. 49, pg. 397.

38. Comparison of Bandwidth - Artificial vs Actual Line Sections - A. VanWeel - Phillips Research Reports - Aug. 50, pg. 241.

39. Artificial Transmission Lines for Video Distribution and Delay - Turner - R.C.A. Review - Dec. 49, pg. 477.

40. L Type Impedance Transforming Circuits - Smith - Electronics - March 42, pg. 48.

41. T and Pi Transformations Simplified - Stockman - Electronics - Oct. 42, pg. 72.

42. Communication Network - Guilleman - John Wiley.

43. Frequency Modulation and Vocal Analysis - Goldman - McGraw Hill.

44. Transients in Linear Systems - Gardner & Barnes - John Wiley.

45. Hyperbolic Chart for $\gamma = \alpha + j\beta$ Conversion - Ware - Electronics - April 45, pg. 148.

46. Modified Resonant Circuits Match Impedances - Sulzer Tele-Tech. - Nov. 50, pg. 41.

47. Bode, H. W. - Network Analysis and Feedback Amplifier Design - D. Van Nostrand .

48. Distributed Amplification - Ginton, Hewlett, Jasberg, Nos - Proc. I.R.E. - Aug. 48, pg. 956 -- Proc. I.R.E. July 50, pg. 748.

49. Traveling Wave Chain Amplifier - Rudenberg Electronics - Dec. 49, pg. 106.

50. Proceedings of the Symposium on Modern Network Synthesis - Brooklyn Polytech. Institute and Office of Naval Research - Sept. 1952.

51. T. V. Master Antenna Systems - Kamen and Dorf. - John F. Rider Publishing Company - 1950.

52. Television Service Manuals - Vols. I thru VII - John F. Rider. - John F. Rider Publishing Company

53. Theory and Design of Television Receivers, Deutach - McGraw Hill 1951.

54. Basic Television, Principles and Servicing - Grob. - McGraw Hill, 1949.

55. Elements of Teleivision Systems, Anner - Prentice Hall, 1951.

56. W. S. Percival, British Patent 460,562 (1935-37).

57. H. A. Wheeler, Wide-Band Amplifiers for Television, Proc. IRE, 27, pg. 437, 1939.

58. Television Antennas for Apartments, Electronics, pg. 100, May 1947.

59. H. G. Rudenberg, Some Characteristics of a Delay-Line Coupled Wide-Band Pulse Amplifier, (Abstract), Phys. Rev. 73, pg. 543, 1948.

60. H. G. Rudenberg, Progress Reports, Cruft Laboratory, Harvard University, Contract N5ori-76 T.O.I.

61. H. E. Kallmann, Television Antenna and R. F. Distribution Systems for Apartment Houses, Proc. IRE, 36, pg. 1153, 1948.

62. J. B. Trevor, Artificial Delay-Line Design, Electronics, pg. 135, June 1945.